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The Proceedings

OF

THE INSTITUTION OF ELECTRICAL ENGINEERS

FOUNDED 1871: INCORPORATED BY ROYAL CHARTER 1921

PART B

RADIO AND ELECTRONIC ENGINEERING
(INCLUDING COMMUNICATION ENGINEERING)

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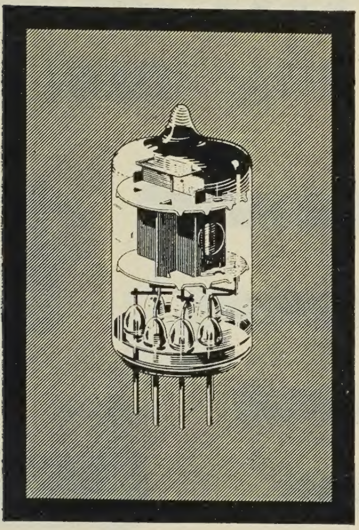
A superior dual control pentode



6AS6

Services type number CV2522

Heater

Vh	6.3	V
Ih	175	mA

Capacitances	With shield	Without shield	
cg1–a	<0.02	<0.025	pF
cg3–a	0.7	0.7	pF
cg1–h+k+g2 +g3+a+s	4.0	3.9	pF
cg3–h+k+g1 +g2+a+s	3.4	3.3	pF
ca–h+k+g1 +g2+g3+s	3.0	2.2	pF
cg1–g3	<0.15	<0.15	pF

Characteristics

Va	120	120	V
Vg3	−3.0	0	V
Vg2	120	120	V
Vg1	−2.0	−2.0	V
Ia	3.6	5.2	mA
Ig2	4.8	3.5	mA
gm (g1–a)	1.85	3.2	mA/V
gm (g3–a)	810	470	μA/V
ra	–	110	kΩ
Vg1 (Ia = 10μA)	–	7.5	V
Vg3 (Ia = 10μA)	−10	–	V

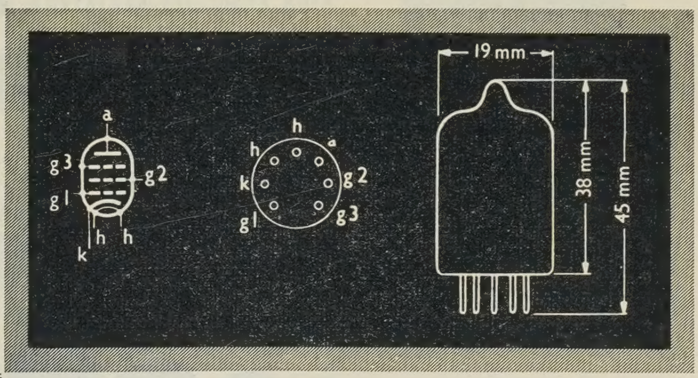
Limiting Values

Va max.	180	V
pa max.	1.7	W
Vg2 max.	140	V
pg2 max.	750	mW
Vg3 max.	27	V
Vh–k max.	90	V

The 6AS6 is a dual control pentode intended for switching or gating applications or for use as a frequency changer.

It is a short suppressor base version of the 6AK5 and is particularly suited for equipment for the American market. Widely used in America and elsewhere, it is manufactured and marketed by Mullard under the American type number 6AS6.

This pentode is designed for operation from 120 volt supplies and is, therefore, convenient for use in conjunction with the 6AK5.



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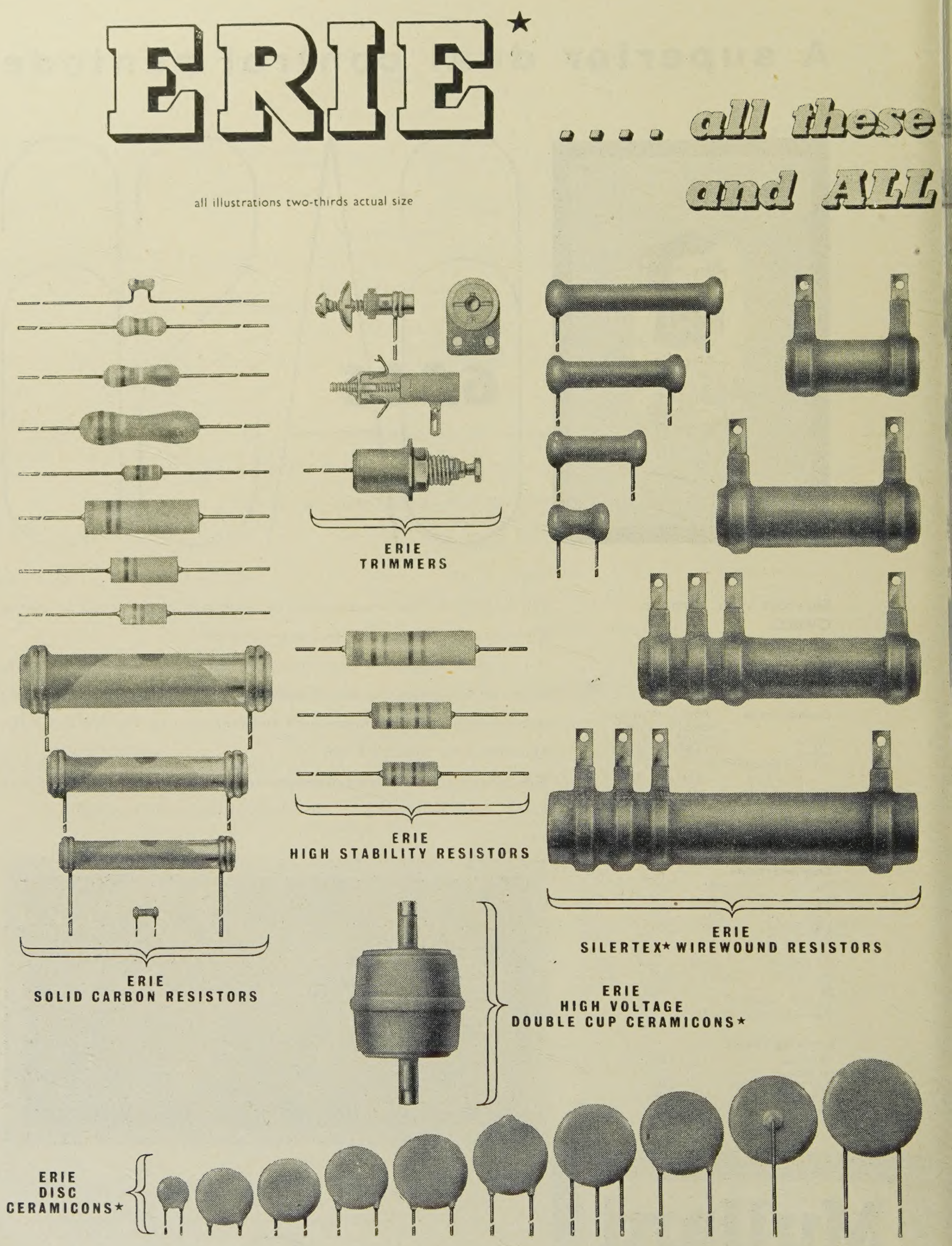
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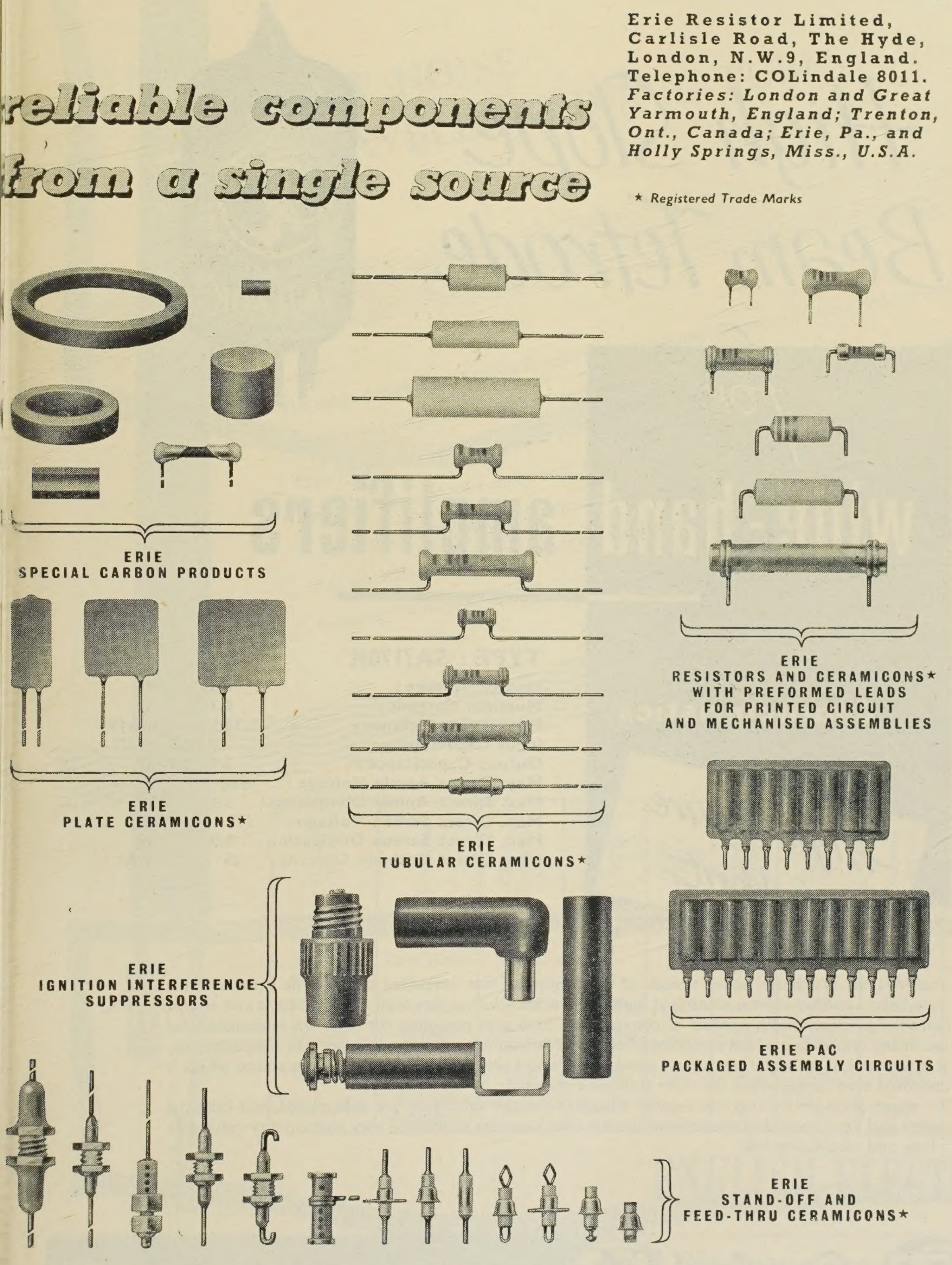
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for wide-band amplifiers

Characteristics

High Figure of Merit

TYPE : 5A/170K

Heater Voltage:	6.3	V
Nominal Current:	0.3	A
Mutual Conductance:	16.5	mA/V
Input Capacitance:	7.9 $\pm$ 0.6	pF
Output Capacitance:	2.9 $\pm$ 0.4	pF
Max. Direct Anode Voltage:	210	V
Max. Direct Anode Dissipation:	3.3	W
Max. Direct Screen Voltage:	175	V
Max. Direct Screen Dissipation:	0.9	W
Max. Direct Cathode Current:	25	mA

The 5A/170K is a beam power tetrode of small physical size developed to meet the demand for a wide-band amplifier valve operating at high frequencies. With a very high figure of merit and almost twice the gain/bandwidth product of conventional high gain pentodes, the 5A/170K is designed for use in any application where a wide-band amplifier is required e.g. radio links, carrier telephony, etc. It is economical on heater power and possesses a remarkably low equivalent noise resistance which is improved even further when the valve is triode connected.

To ensure good electrical contact under all conditions the valve pins are gold plated, and both the design and rigid control of manufacturing processes combine to make a very high quality valve with a long and trouble-free life.

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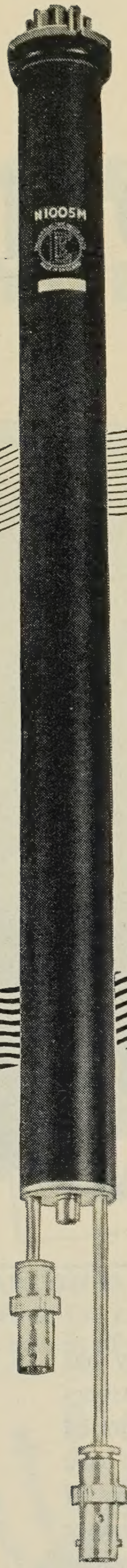

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The range designed by English Electric Valve Company includes low noise, voltage amplifier and power tubes, with outputs from 1mW to 16W.

Type N1005M illustrated is a low noise tube specially designed to operate over a frequency range of 3600-4200 Mc/s. It permits the use of r.f. amplification in radar, tropospheric scatter and other microwave equipment.

Full particulars of this tube and other units specially designed for use in the higher frequency bands are available on application.

E.E.V. Type	Function	Centre Frequency (Mc/s)	Maximum Output	Noise Factor (dB)	Gain (dB)	Helix Volts	Collector Current	Focusing Field (Gauss)
N.1001	Power	2000	16W	–	26	2600	40mA	600
N.1002	Low Noise	2000	1mW	10	24	550	200μA	200
N.1004	Power	4000	4W	–	21	2600	20mA	450
N.1005M	Low Noise	4000	1mW	11	22	360	200μA	350
N.1013	Voltage Amplifier	2000	200mW	20	32	650	4mA	300
N.1017M	Low Noise	1200	1mW	10	20	700	200μA	200
N.1018M	Voltage Amplifier	4000	100mW	20	30	630	2mA	350

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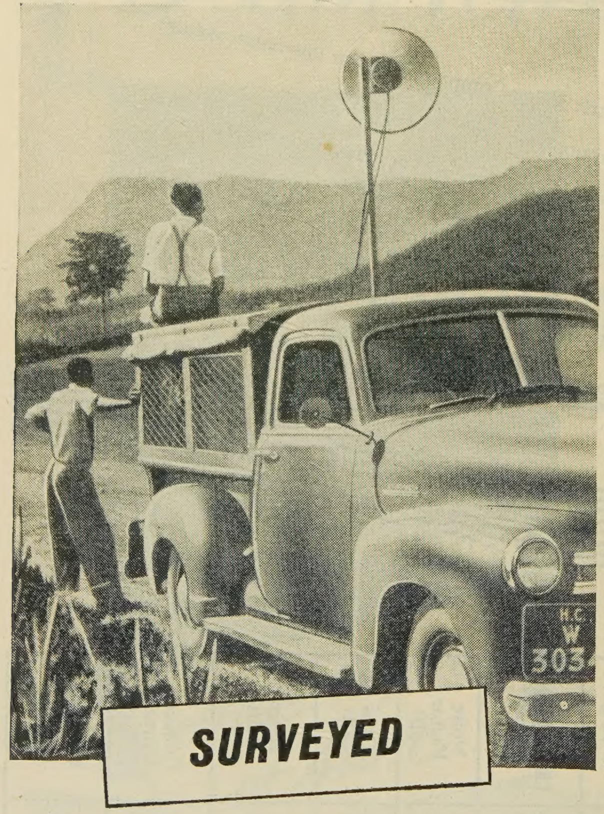

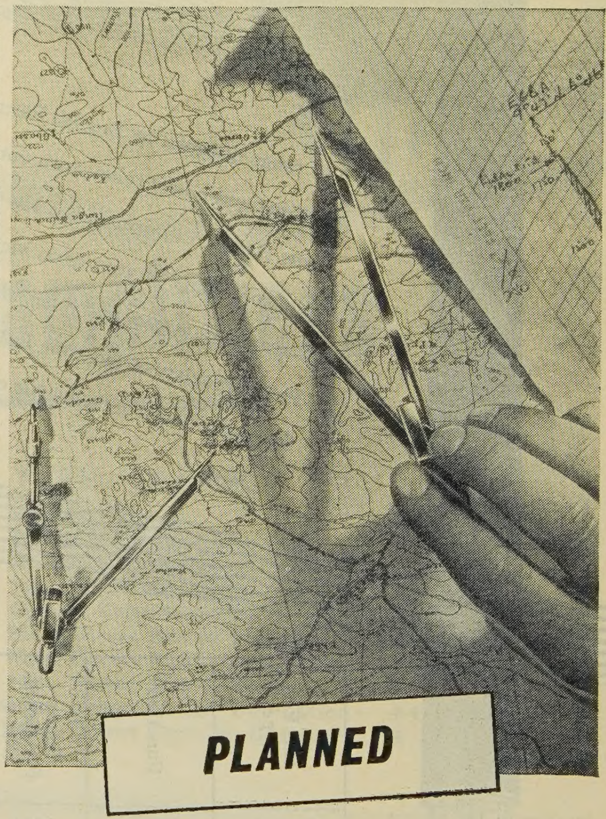

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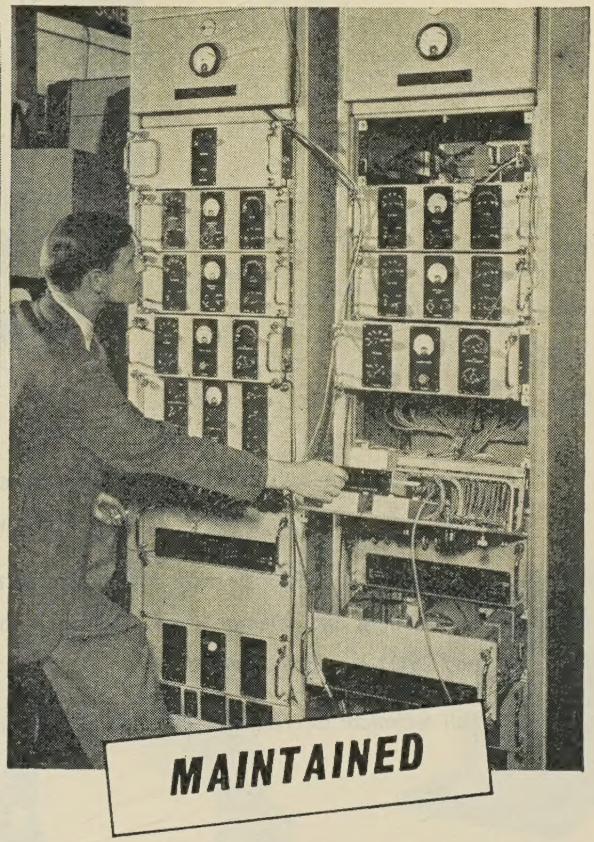

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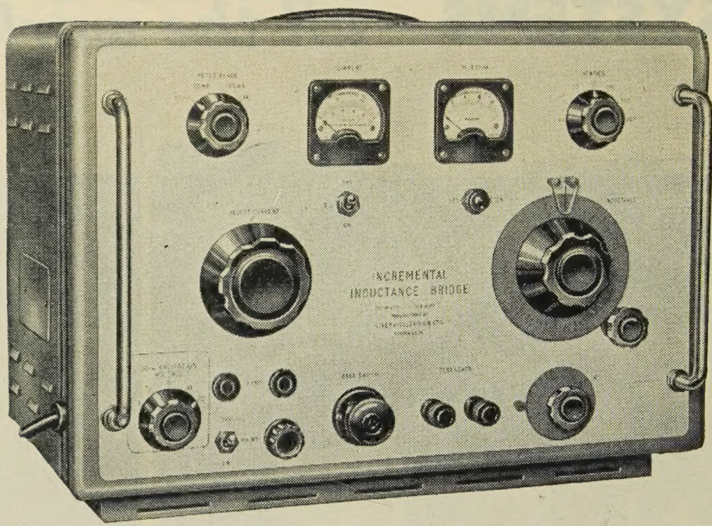

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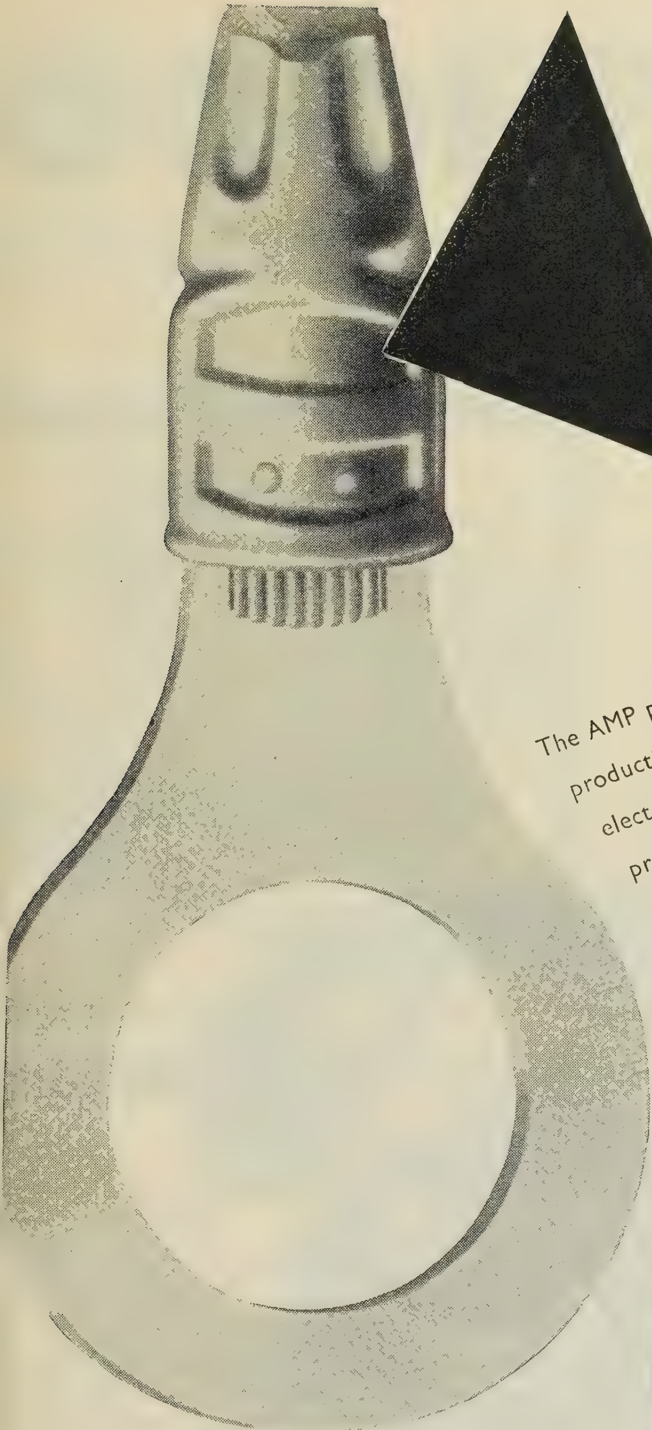

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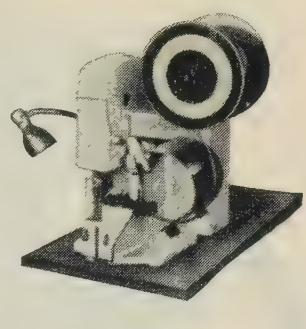

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PERFORMANCE

The rocking-armature receiver has an excellent frequency response and an output 10 dB higher than its predecessors. Part of this improvement is transferred by the induction coil to the sending side, giving a resultant improvement in performance of 6 dB on sending and 4 dB on receiving.

This improved performance means that the new G.E.C. telephones will operate over a local-line loop of 1120 ohms of $6\frac{1}{2}$ lb cable (·5 mm conductor) with a performance equal to that of previous telephones operating over a local-line loop of 660 ohms, i.e. a local line may be extended by 70%. Alternatively, the new telephones will operate over a length of 4 lb cable (·4 mm) with a performance equal to that of a previous telephone operating over the same length of 10 lb cable (·6 mm.).

Despite the increased level of performance, the maximum amount of sidetone suppression has been retained.

COLOUR

In addition to the normal black instrument, a range of two-tone telephones can be supplied in which the case is coloured and all other parts, including the dial, are black. The range of colours is red, green and ivory. The two-tone telephones have the following advantages over the all-colour telephones:—

★ **More delicate shades can be used since shade matching is eliminated.**

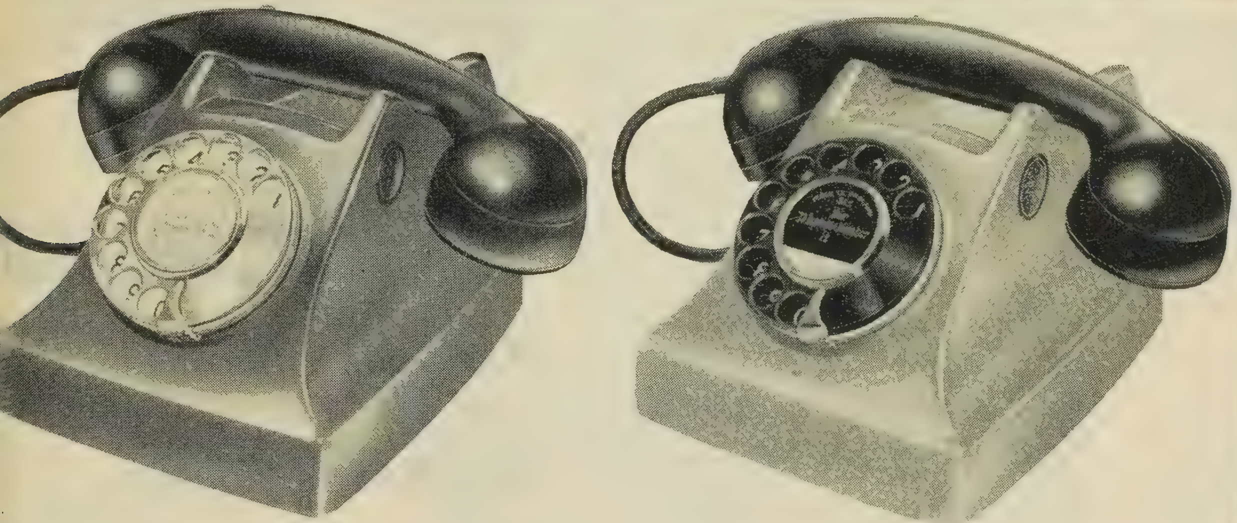
★ **The number of spare parts required to be held by an Administration using more than one colour of instrument is greatly reduced.**

The colour

THE GENERAL ELECTRIC COMPANY LIMITED OF ENGLAND

TELEPHONE

The appearance has been redesigned to blend with modern styles of decoration, whilst retaining the dignity essential to office furnishings.



of a telephone can be changed simply by changing the case

APPEARANCE

The pleasing appearance is achieved by blending a number of curved lines and surfaces to form outlines for the case and handset that are in complete harmony with one another. The camber of the sloping front houses the dial in the automatic telephone and the dial dummy in the C.B. set. The incline of the telephone front is such that the telephone is easy to use, and the dial numbers easy to see, whether the user is in a standing or sitting position.

A curved insert is fitted in each side of the case in front of the cradle to provide finger-tip grips for lifting and carrying the instrument. When resting in the cradle, the curved handset gives the telephone a domed silhouette in accord with the remainder of the instrument. The increased curvature of the handset over previous types gives greater comfort to the user, and tilts the transmitter to a more sensitive position.

TROPICALISATION

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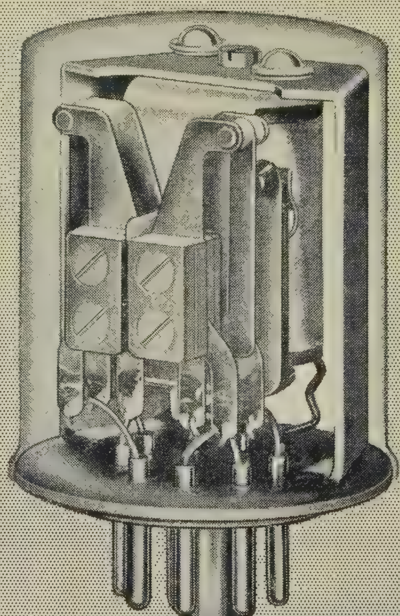
FIRST

TRANSISTORISED RELAY

The Hermetically Sealed 595HS



★ The 595HS can be controlled by ultra sensitive contacts handling 0.4mA. at 2V. Contacts will handle 5A. at 230V. A.C.

★ The 595HS is made to withstand exceptionally heavy shock and vibration.

★ The 595HS is made to withstand dirt and humidity indefinitely.

★ The 595HS can be obtained with various contact assemblies.

★ The 595HS is low in price because of its novel design.

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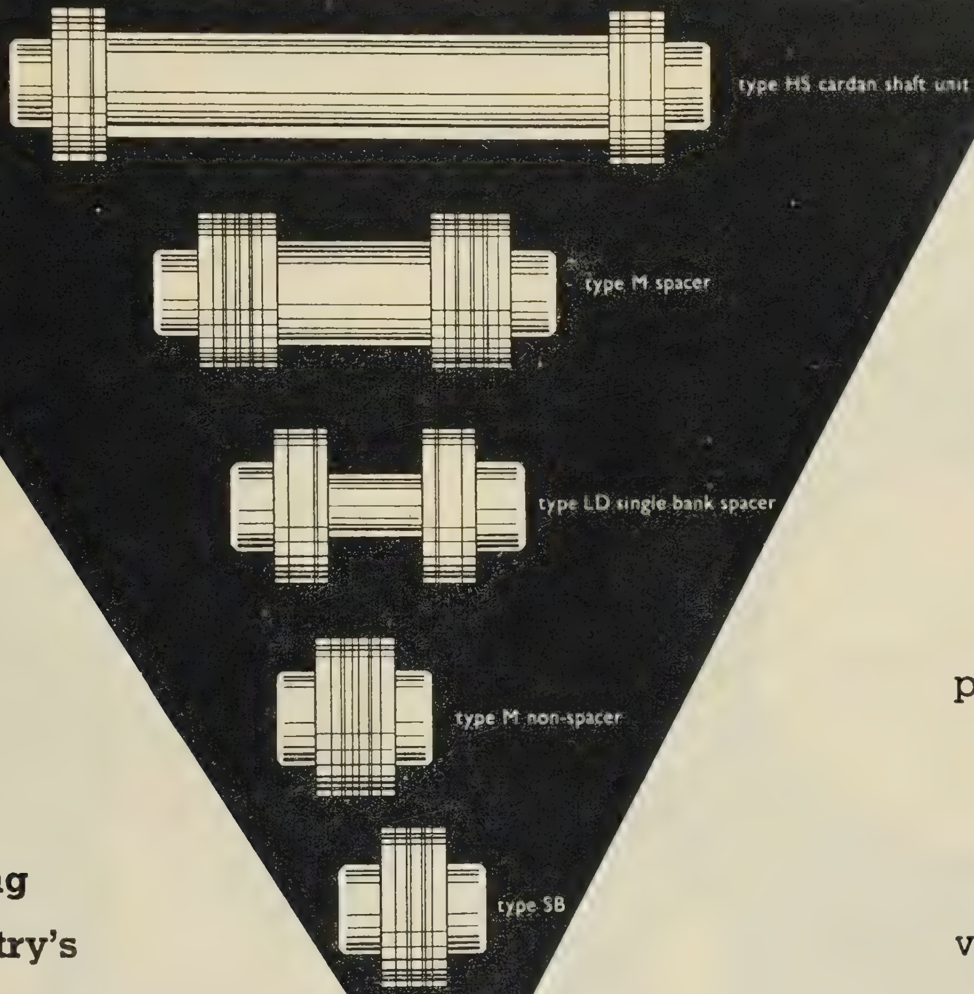

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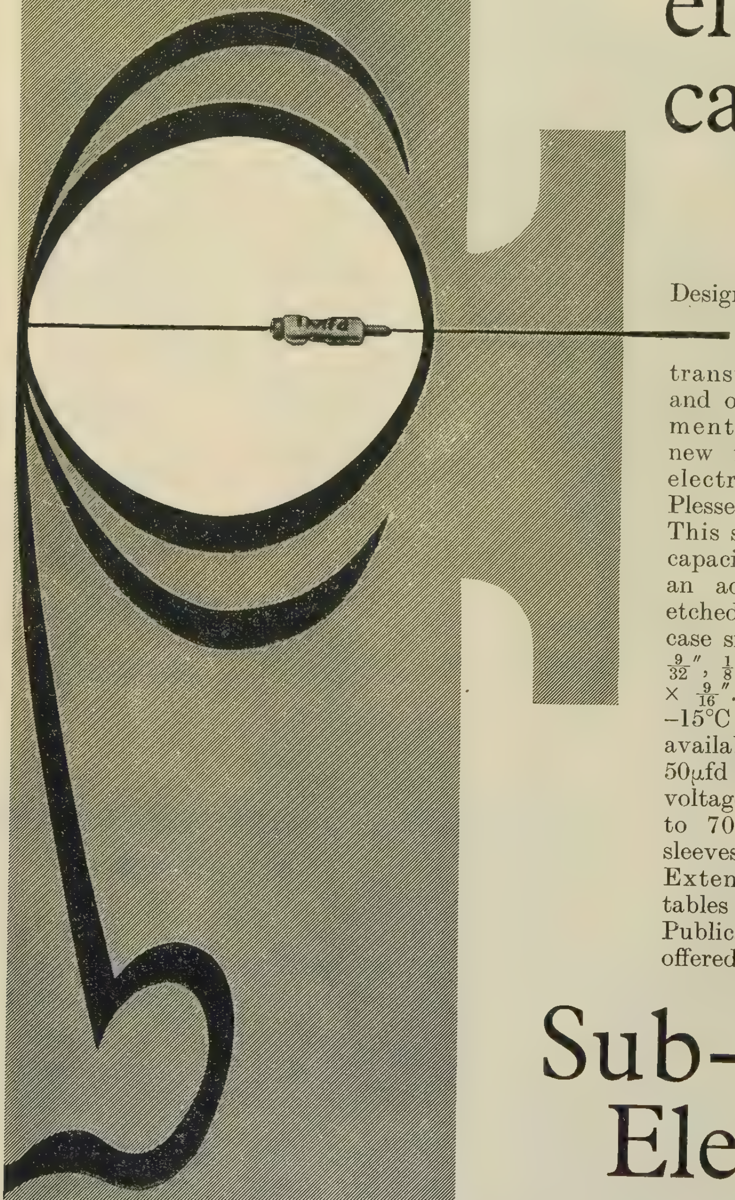

Design Engineers concerned with transistor circuitry for hearing aids, midget transmitters and receivers and other miniaturised equipments, will welcome this new range of sub-miniature electrolytic capacitors by Plessey.

This superior all-aluminium capacitor is made possible by an advanced application of etched foil construction. Four case sizes are available; $0.1'' \times \frac{9}{32}''$, $\frac{1}{8}'' \times \frac{7}{16}''$, $\frac{3}{16}'' \times \frac{1}{2}''$, and $\frac{1}{4}'' \times \frac{9}{16}''$. Temperature range is $-15°C$ to $+60°C$. Capacities available are from 0.5μfd to 50μfd according to working voltages which range from 1.5v to 70v. Plastic insulating sleeves can be fitted if required. Extensive details and data tables are set out in Plessey Publication No. 847 which is offered on request.

Sub-miniature Electrolytic Capacitors by **Plessey**

COMPONENTS GROUP · SWINDON COMPONENTS DIVISION
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SC2b

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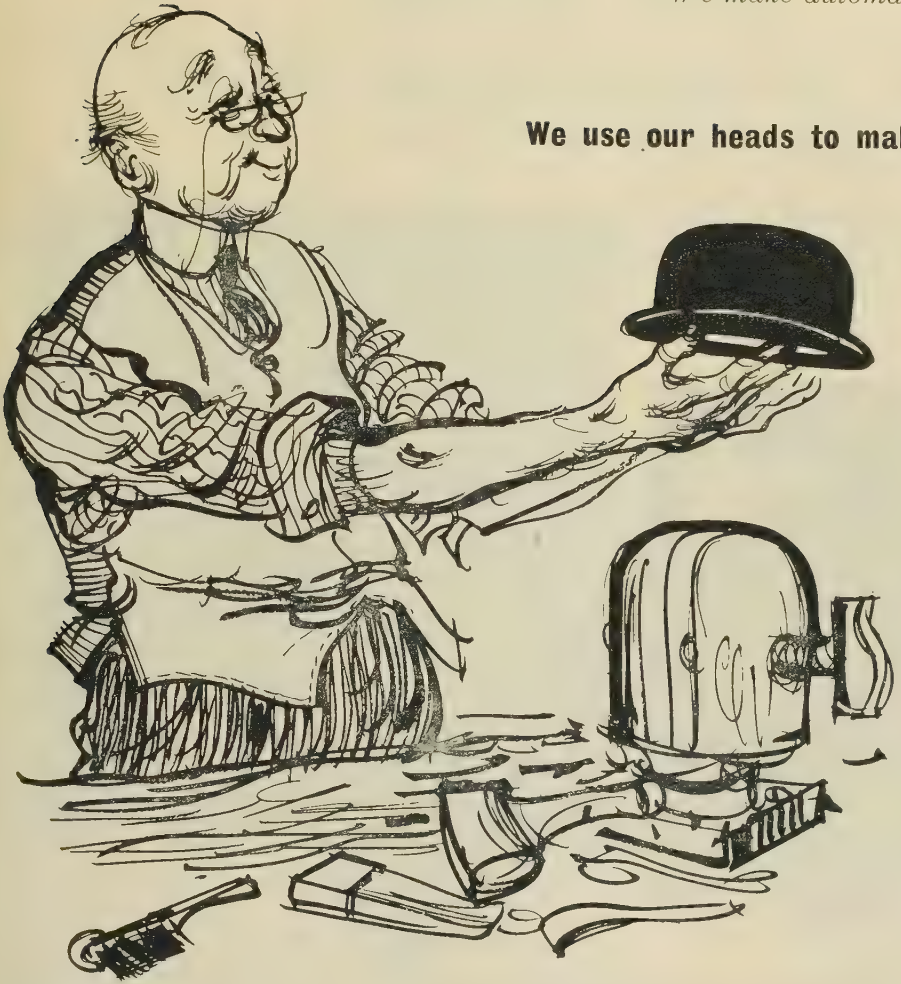
We use our heads to make it fit



There is standardised Austinlite Generating Plant that fits many jobs perfectly—even difficult and unusual ones. But the idea behind Austinlite is different. It is the idea of an uninterrupted supply of electricity under conditions which may be so difficult that no standard plant could be expected to do the job at all. It is the skill and experience of engineers who have provided such a supply in many parts of the World that we have to offer—rather than various arrangements of engine, generator and bedplate.

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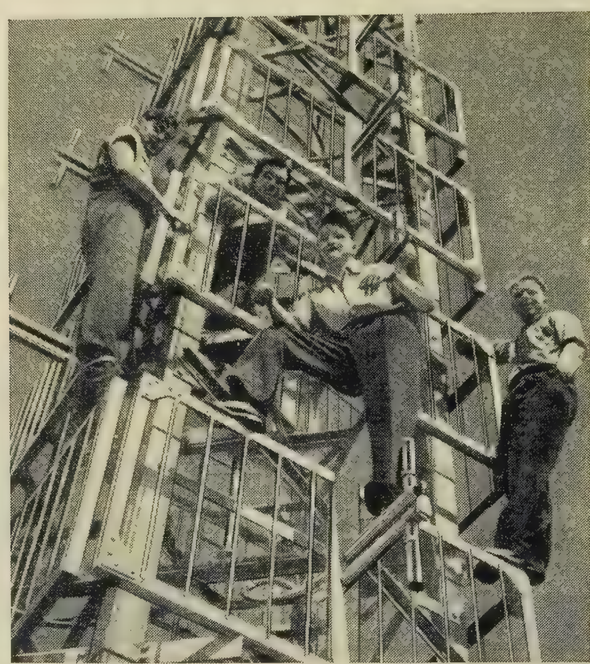
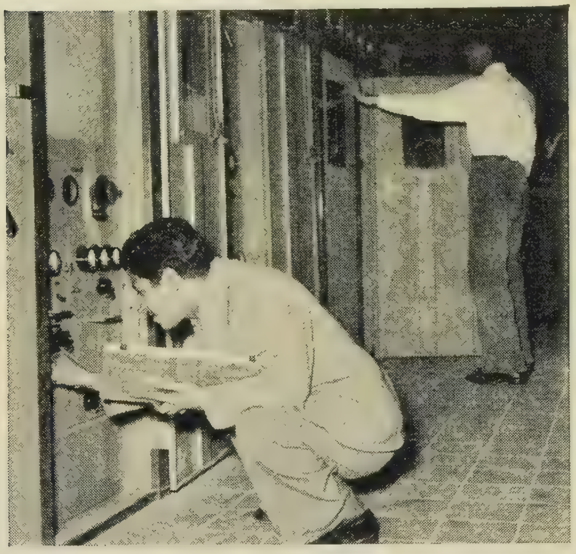
From on the map to on the air





through Marconi's experienced hands

Broadcasting and television authorities all over the world look to Marconi's for much more than the supply of equipment. The company has been called on for every aspect of the provision of a broadcasting service, from the survey of propagation problems in the area to be served, through the complete building of the transmitter stations and the installation of the programme input equipment, to the erection of the aerials, maintenance, and the training of technical staff. No other company in the world tackles such matters with the experience, research facilities, skill and resourcefulness of Marconi's.

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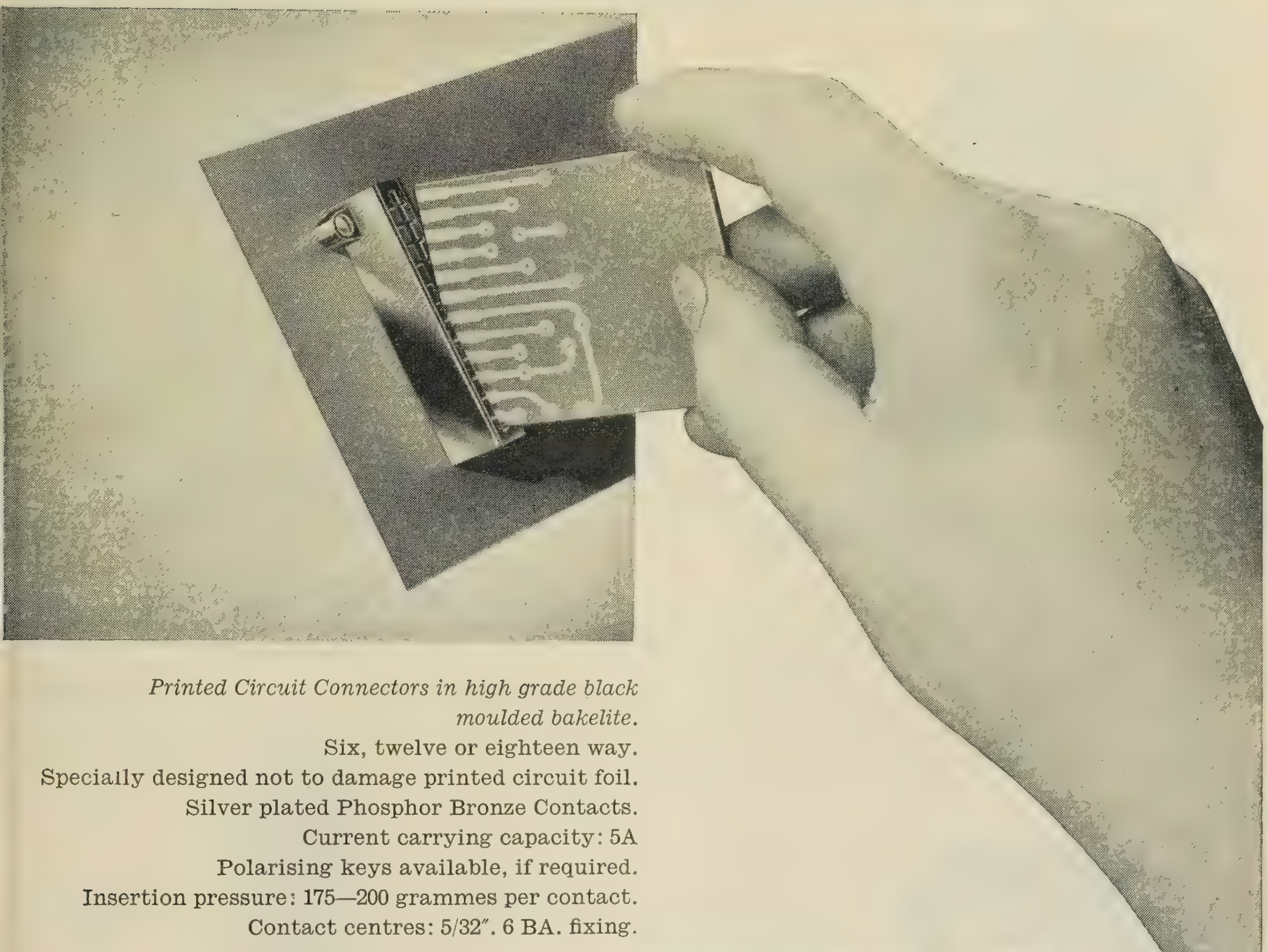
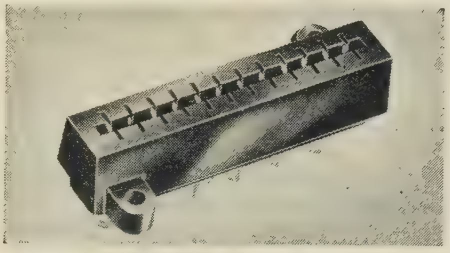
LB 6

PRINTED CIRCUIT CONNECTORS



Printed Circuit Connectors in high grade black moulded bakelite.
Six, twelve or eighteen way.
Specially designed not to damage printed circuit foil.
Silver plated Phosphor Bronze Contacts.
Current carrying capacity: 5A
Polarising keys available, if required.
Insertion pressure: 175—200 grammes per contact.
Contact centres: 5/32″. 6 BA. fixing.



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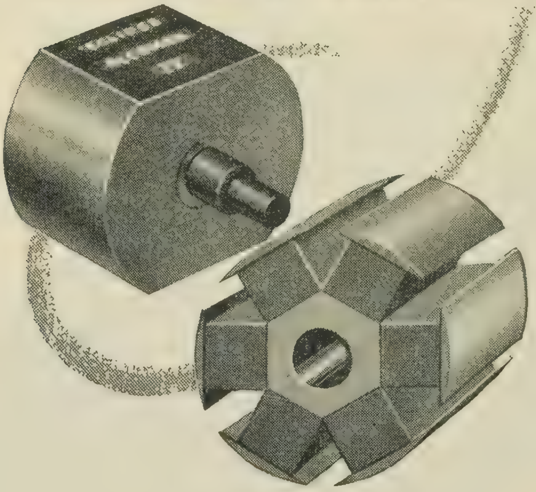
Why Alcomax IV FOR ROTATING MAGNETS?



Development of very high coercivities generally necessitates some sacrifice of energy content, but in Alcomax IV a material is available with energy content only slightly less than that of Alcomax III and with a still higher coercivity. Alcomax IV is outstanding in having these two qualities simultaneously. It is particularly advantageous for very short magnets, in systems requiring a high flux density in a long gap, and in rotating machines. Ask for Publication P.M. 131/53 "Design and Application of Permanent Magnets."

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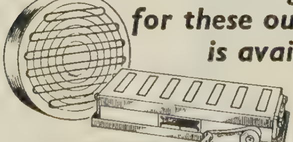

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M5

WHY IT PAYS TO USE
Ersin Multicore Solder

Leading manufacturers prefer Multicore Solders. Many in this country are changing to the new SAVBIT Type 1 Alloy. This alloy was specially developed to reduce absorption of copper into the alloy — the main cause of bit wear. In fact, the life of solder bits can be extended by up to 10 times, which represents a considerable saving in replacement costs.

- Ersin Multicore is the only solder containing 5 cores of Ersin Flux, a high grade rosin which has been subjected to a complex chemical process to increase its fluxing action, whilst still retaining the non-corrosive properties. Both the standard alloys and the new Savbit alloy incorporate Ersin Flux which prevents formation of oxides during the soldering process and also removes any oxide layer on the metal.
- Five cores of Flux ensure flux continuity throughout the length of the solder wire — there are no lengths without flux.
- The correct proportions of flux to solder are always assured — no extra flux is required. Five cores of flux provide thinner solder walls, giving instantaneous melting.
- Soldered joints made with Ersin Flux do not corrode even after prolonged exposure to any degree of humidity.
- Only the finest virgin tin and lead are used in the manufacture of Ersin Multicore.

FOR FACTORY USE. The economies effected by using Ersin Multicore Solder play an important part in cutting production costs and keeping down the price of equipment. You get more joints per lb. of Ersin Multicore—there is no waste. Soldering with Ersin Multicore is quicker too and every joint is a perfect electrical connection. Ersin Multicore Solder is made as standard for factory use in 6 alloys and 9 gauges; Savbit alloy is available in 3 gauges. Both are supplied on 1 lb. and 7 lb. reels. Bulk prices on application.
TECHNICAL INFORMATION. Electrical engineers and technicians are invited to write for comprehensive technical literature about Ersin Multicore Solder containing useful tables of melting points etc., and samples of alloys.

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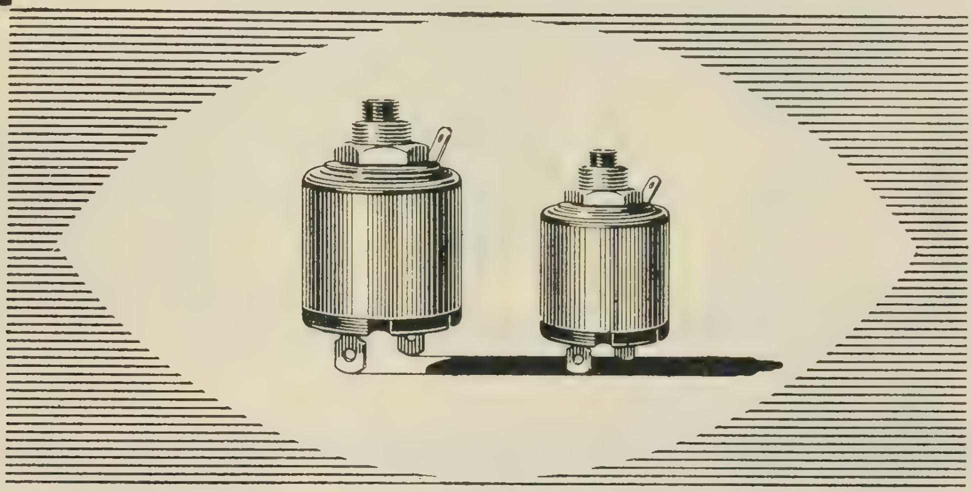
CAN NOW BE ADJUSTED . . . AFTER MOUNTING
OVER AN AVERAGE RANGE OF 10%
WITH A SETTING ACCURACY OF ·5%



These new Mullard 14mm and 18mm pot cores are completely self contained, simple to mount and easily adjusted after mounting. Unique features include single hole fixing and two or four way terminal plates. Adjustments of inductance are made by means of a screw which varies the position of a magnetic shunt in the centre of the core; in many cases this eliminates the need for trimming capacitors. Designers will see from the brief characteristics listed here that Mullard adjustable pot cores are particularly suitable for use in high grade communications equipment, tuned circuits and filter networks. Those requiring further technical details are invited to write to the address below.

14mm Pot cores

Four types available	LA32 — 35
Air gaps	From 0.2mm — 0.5mm
Frequency range	10 Kc/s — 100 Kc/s
Q valves in the higher frequency range	> 200

18mm Pot cores

Four types available	LA42 — 45
Air gaps	From 0.3mm — 1.0mm
Frequency range	10 Kc/s — 100 Kc/s
Q values in the higher frequency range	> 300

Mullard

'Ticonal' permanent magnets
Magnadur ceramic magnets
Ferroxcube magnetic cores

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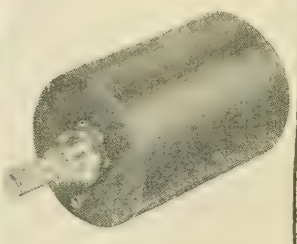
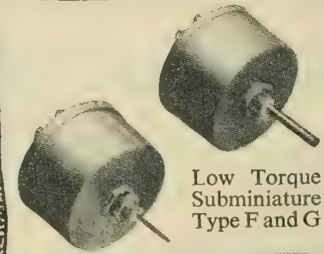
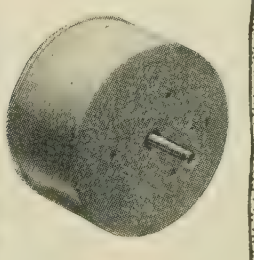
by P·X·F



Helical Potentiometer Type PX 4/H.10. Turnings: 3 and 10.



Low Torque Subminiature Type F and G



Toroidal Potentiometer Type B and E with ball races.

Ceramic Potentiometer 8 sizes 10–1,000 watts.

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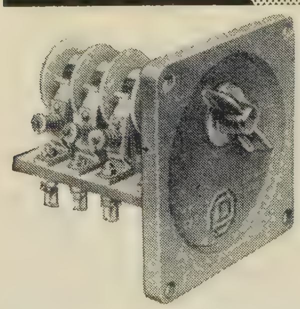
A SELECTION FROM

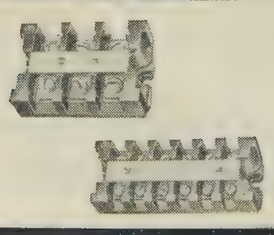
Miniature plunger, door interlock switch with N.O. contacts, 2-amps. 550-volts A.C. Size $2\frac{1}{8}''$ long × 1″ dia. (Type C 32 L 310).

A LARGE RANGE OF



Flush-mounting T.P. Isolating Switch for building-in Machine Tools, etc. 30-amp. 550-volts (Type C 95).

CONTROL ACCESSORIES



Terminal blocks (Type J 96) with clamp type terminals, white marker strip. Generous clearances between phases and to earth.
Made in the following sizes: 15-amp. 550-volt. 3-, 4- and 6-way. 30-amp. 550-volt 3- and 4-way.

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Available 2-, 4- or 8-poles (with one or two coil circuit change-over contacts), fine silver double-break main contacts rated at 10-amps. 550-volts, Any pole can be N.O. or N.C.

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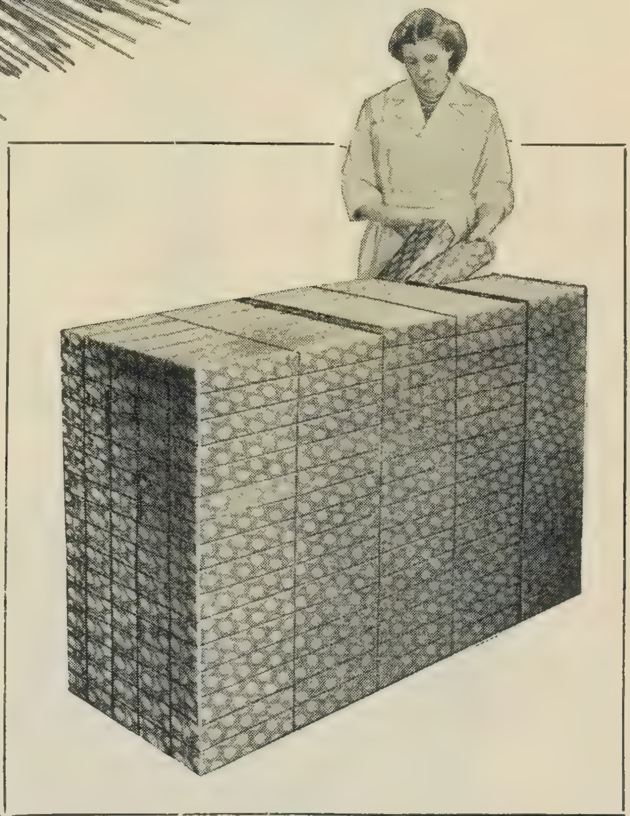
box clever

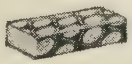
with four new autopacks

for RESISTORS

DUBILIER AUTOPACKS—designed primarily for loading hoppers for automatic feed systems—also solve the spiky problem of resistor storage. The resistors* are housed in packs of 50, 100, 500 and 2,500 for type BTS, and in packs of 100 and 500 for type BTB—all lined up with connecting wires dead straight, ready for immediate use—and taking up very little space in the process. The large *and* the small user would be well advised to take stock of these handy, space and time-saving packs.



50



100

500

2500

No storage problems here!

This modest sized stack contains no fewer than one million separate resistors. A pack of 50 would occupy no more room than 10 cigarettes on the workbench. Whether you use resistors in tens, hundreds or thousands—see Dubilier about these new space-saving Autopacks today.

**The resistors are available in two ratings: BTS ½ watt, BTB 1 watt at 70°C. Resistance range is 100 ohms to 10 megohms (BTS) and 390 ohms to 22 megohms (BTB). Each type is completely protected by a phenolic resin housing which is sealed at the ends.*

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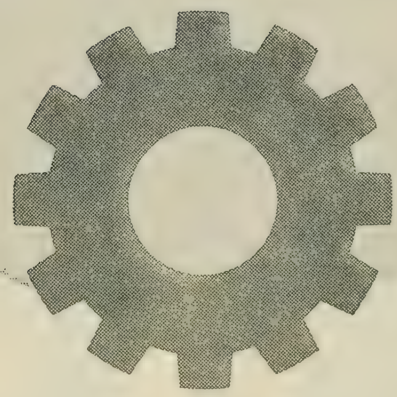
DN 173

PHILIPS



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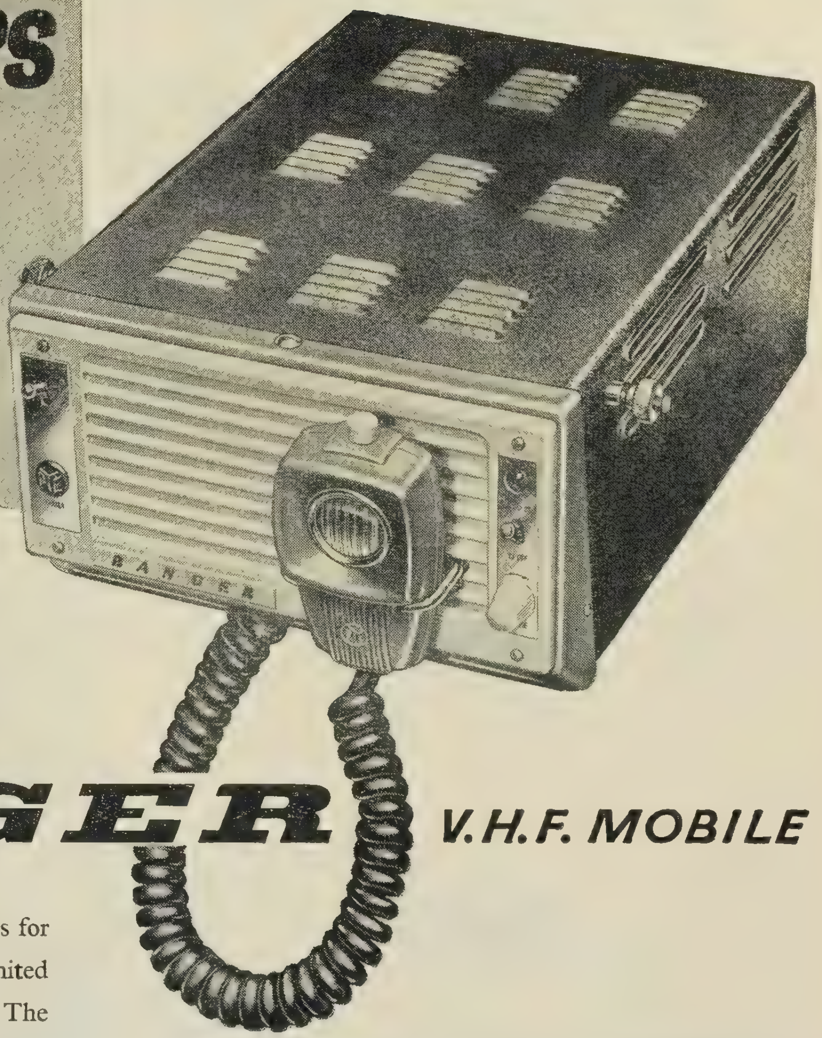

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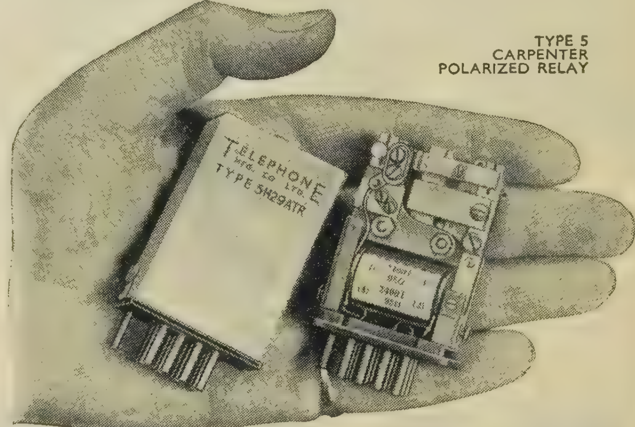

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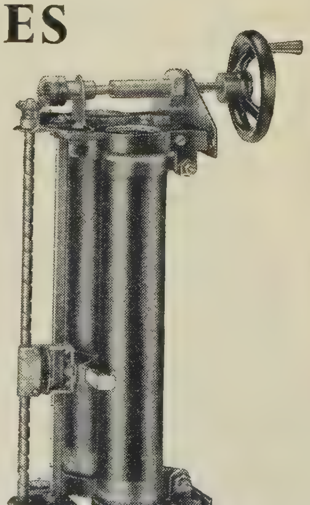

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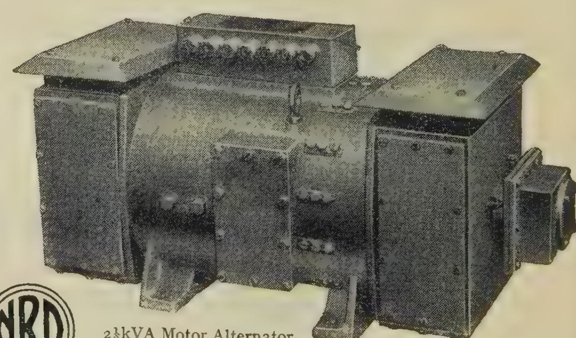
Also makers of Rotary Transformers and Converters, Wind and Engine-Driven Aircraft Generators, High Tension D.C. Generators, and Automatic Carbon Pile Voltage Regulators.



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- Low Weight
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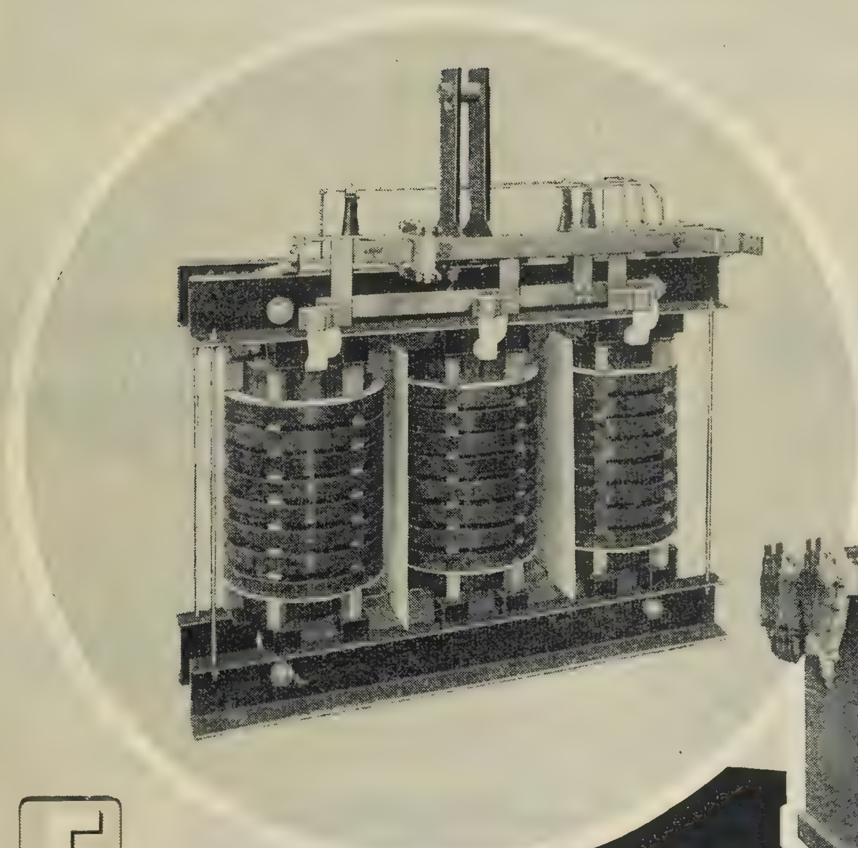
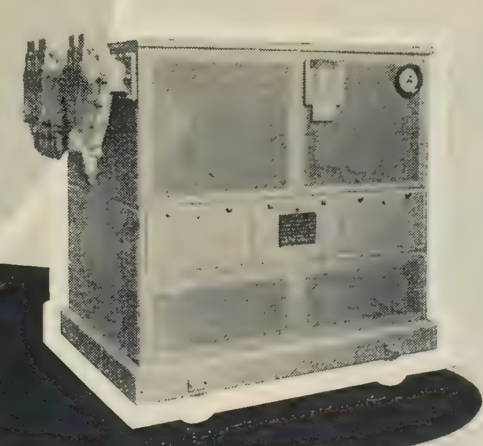


Illustrated is a 500 kVA, 3-phase, 11,000 volt Ferranti Distribution Transformer using 'Class H' Insulation, as supplied to the Air Ministry.



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The meter will differentiate between A.C. and D.C. supply, the switching being electrically interlocked. The total resistance of the meter is 500,000 ohms.

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VOLTAGE: A.C. and D.C. 0 to 1,000 volts
RESISTANCE: Up to 40 megohms.
CAPACITY: .01 to 20μF.
AUDIO-FREQUENCY POWER OUTPUT: 0–2 watts.
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The instrument is self-contained, compact and portable, simple to operate and almost impossible to damage electrically. It is protected by an automatic cut-out against damage through severe overload.

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Weight: $6\frac{3}{4}$ lbs.

List Price:
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They are used
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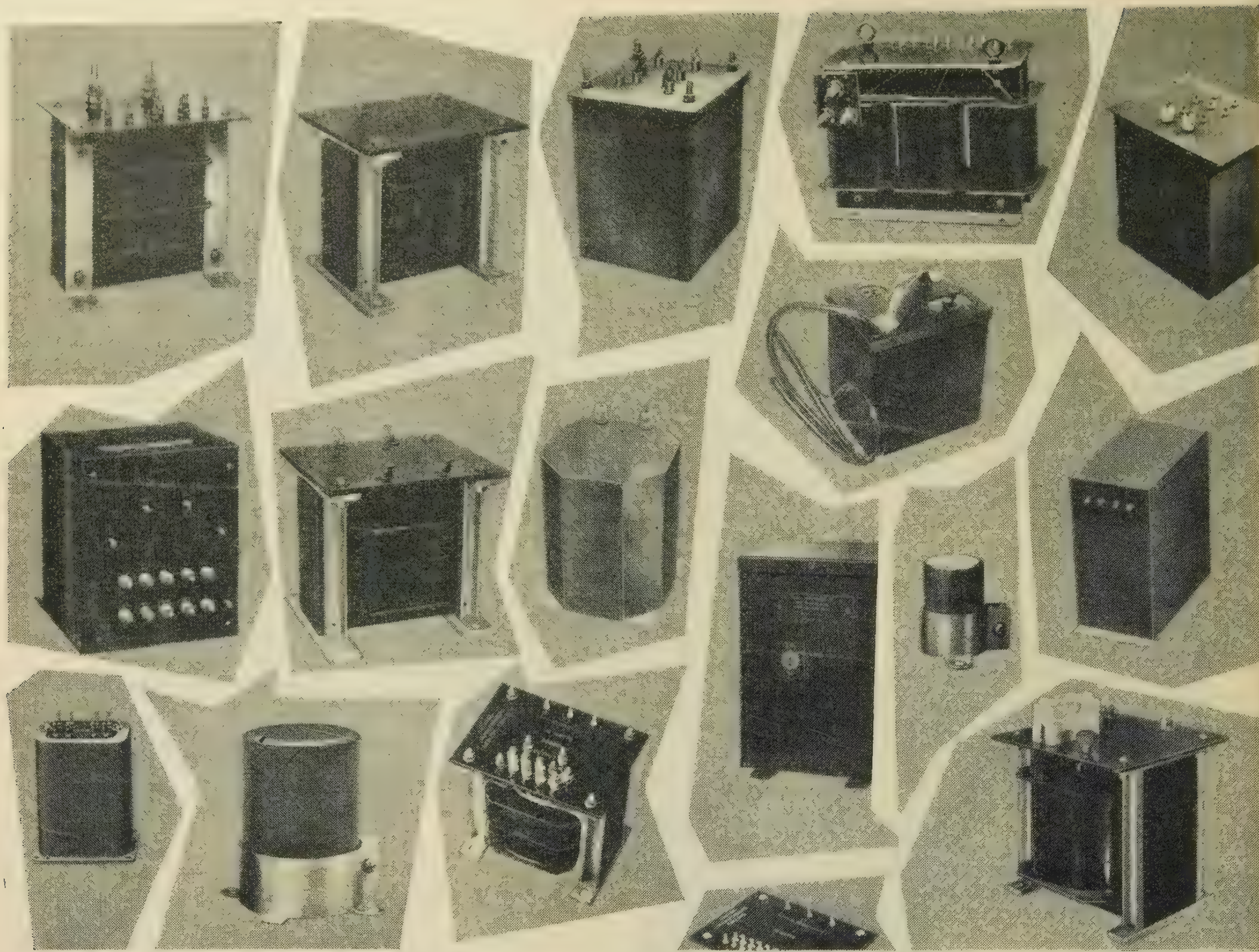

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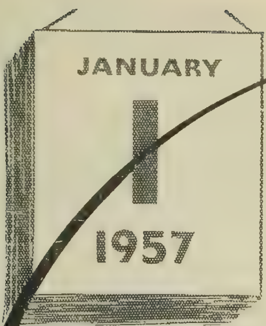

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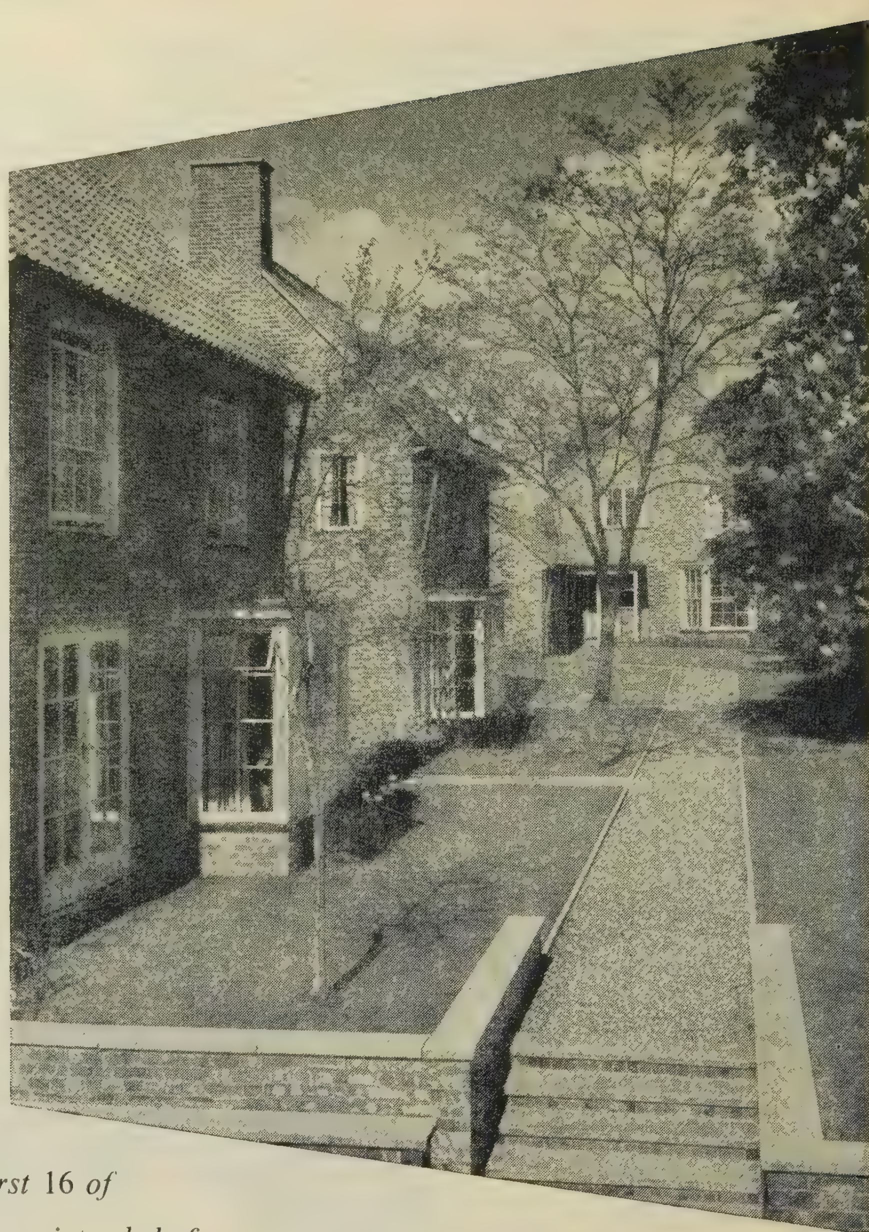

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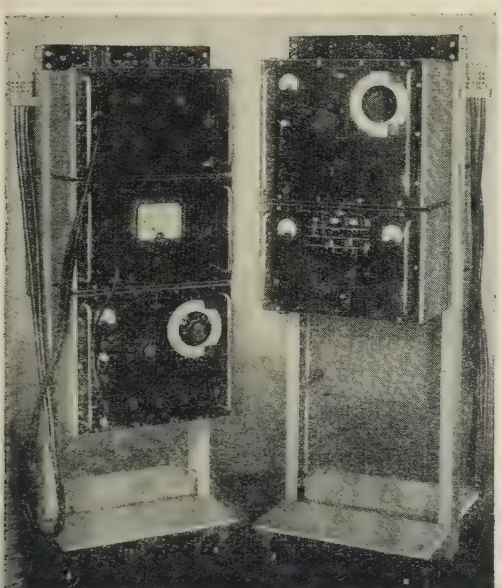

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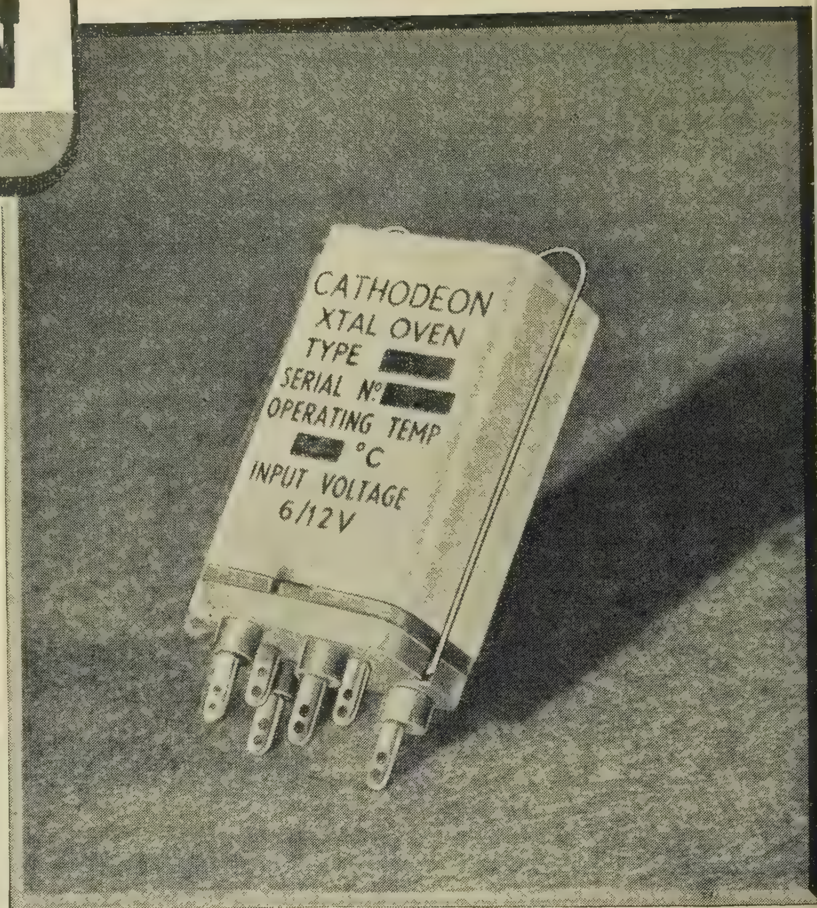

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THE PROCEEDINGS OF THE INSTITUTION OF ELECTRICAL ENGINEERS

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Oct. 1956

INAUGURAL ADDRESS

By Sir GORDON RADLEY, K.C.B., C.B.E., Ph.D.(Eng.), President.

(*Address delivered before* THE INSTITUTION *4th October*, 1956.)

I would, first of all, express my appreciation of the signal honour which my fellow-members have paid me in electing me as President of The Institution. The honour is one which gives me the greater pleasure because my responsibilities are no longer entirely engineering. A high tradition of service has been established by those who have filled the presidential chair and, in my turn, I will do my utmost to serve my fellow-members to the best of my ability, and to the extent that health and strength and the obligations of being a public servant permit.

The Institution

Last year Sir George Nelson reminded us of Charles William Siemens, the first President of The Institution, or of the Society of Telegraph Engineers as it was then called. To-night I recall the second President. Frank Ives Scudamore, President in 1873, was Second Secretary to the Post Office. In his Inaugural Address he said he was 'not merely thinking of the benefits which the leading engineers of the Postal Telegraph Department derived from contact in this room with the engineers of other services'. He was looking beyond, but only at that stage to the 'great body of persons' who were then engaged in the practice of telegraphy throughout the country. During the past 80 years the uses of electricity have grown from that simple start until they have become essential to every phase of contemporary existence, social, business and national defence. The Institution has adapted itself to this growth—first by change of name until it acquired its present title representative of the whole profession in 1888, second by formation of the Specialized Sections starting with the Wireless Section in 1919.

The branch of electrical engineering in which I have worked—telecommunication—has much in common with physics, whereas those members who are, for example, engaged in the building of electricity generating stations have a traditional affinity with civil and mechanical engineering, which are the classical fields. The opinion has been expressed from time to time that the organization of The Institution does not sufficiently recognize this difference in outlook. But, especially at a time such as this of rapid advance, there can be no hard and fast boundaries between the so-called light- and heavy-current technologies. Much that we do requires both. On the other hand, the present organization will not remain adequate indefinitely. The last Annual Report did well to remind us that our organization must remain flexible and The Institution must, 'as a living organism continually re-adapt itself to the requirements of the time'.

The Institution's approach to developments which lie partly outside its own domain must also be experimental. The setting up of the British Nuclear Energy Conference is an illustration. Here we have had the major professional institutions combining to create a forum in which the basic sciences and established technologies can contribute to a new advance.

Before passing to the main theme of my Address, I would add one comment on the place of the electrical engineer in society. There has been a growing realization that national survival in war and standards of living in peace alike depend on the results of scientific research, and, in consequence, the scientist has won his way into the inner councils of the nation. But in public administration or in industry there is just as great a need for the engineer as for the scientist. It is the engineer who alone can bring to the determination of policy the training and experience necessary to translate scientific ideas into machines.

The Institution, recognizing the importance of the professional electrical engineer in our national life, has had careful regard to the standards of professional qualification. These have been adjusted from time to time to take into account the more exacting nature of the service that the engineer is now expected to give, whether it be in power generation and distribution, radio communication, or the many applications of electronics. In any one of these fields there has been a significant increase in the amount of theoretical knowledge to be possessed. At the same time, rapidly changing technology has in no way diminished the need for practical training and experience. Born in the lecture theatre, the engineer is only brought up in the workshop.

Inland Telecommunication

Those of my predecessors, as Engineers-in-Chief of the Post Office who have also had the honour of being Presidents of The Institution, have mostly taken advantage of evenings such as this to describe to their fellow-members telephone development within the United Kingdom. I want to pass later to a wider theme of world telecommunication, but I must refer first to the inland system, for which the Post Office has a two-stage plan. The first stage is to install enough additional plant, chiefly local cables and exchange equipment, to satisfy the outstanding demand for telephone service. Considerable progress has been made, and

when the seven millionth telephone was connected in July the size of the system had been practically doubled since the war. The second stage comprises the progressive mechanization of the system with the introduction of new facilities.

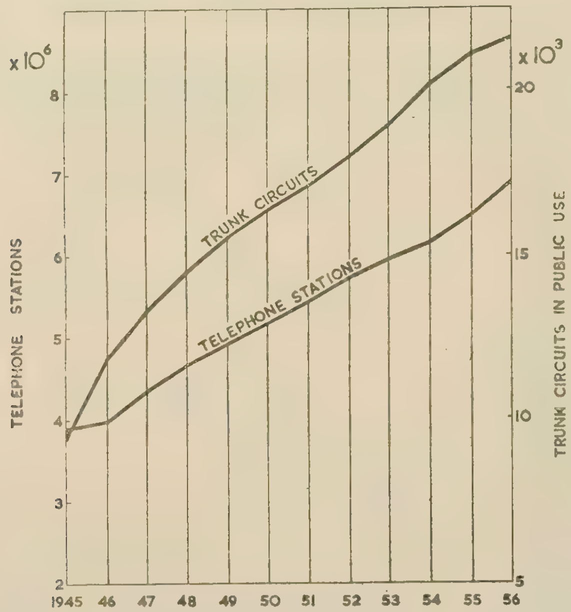


Fig. 1.—Post-war growth of the telephone system.
Figures for each year are as at 31st March.

Fig. 1 shows the post-war growth of the telephone system as a whole, and of the long-distance facilities in particular.

During the past 25 years the results of scientific research have been apparent in building up the facilities for long-distance communication. Cables transmitting 600 speech channels within a single coaxial tube have been installed between the main centres of population. The same line plant, with appropriate repeaters to transmit a slightly wider band of frequencies, is used to provide television links rented to the broadcasting authorities. With little modification it could cater for 1 000 telephone channels on each tube.

As traffic demands and opportunity offers it is proposed to reduce the spacing between repeater stations on some routes to about three miles. Modified in this way, existing cables will cater for about 1 000 telephone channels plus a 405-line television channel on each tube. Alternatively, the tube could carry 2 000 telephone channels, but it is doubtful whether the risk of losing such a large number due to a single line fault would make this arrangement attractive for general adoption. The ability to transmit telephony and television alternatively, or simultaneously, means that the repeaters must meet the requirements both of telephony in respect of low intermodulation, and of television in respect of minimum phase distortion. This is not easy. Development in any case depends on the use of valves with a performance superior to that of any at present in use in this country.

In the first Elizabethan Age, beacons signalling from hill-top to hill-top were used to transmit simple messages in emergency. The modern equivalent is the radio relay system. The facilities provided are much greater—as are the difficulties to be overcome before construction can begin. In order to build a radio station on top of a hill in rural England it is necessary to obtain the consents of no less than 25 authorities. A radio relay system has stations at intervals of between 20 and 50 miles, each station having a line of sight to its neighbours, and operates on carrier frequencies above 1 000 Mc/s. A microwave relay system of this kind is already in extensive operation in the United States, and H. Faulkner planned a comparable system extending from south to north through the United Kingdom. Essentially the system will carry up to six independent radio transmissions in each direction, and each transmission will be capable of bearing up to 10 supergroups (600 channels) of telephony, or one television channel. At terminal and intermediate stations the separate transmissions will be handled in independent amplifying equipments. This will make it convenient to lead off transmissions, as required, by spurs to cities on either side of the route. Valuable economies can, however, be achieved by the use of common aerial and waveguide systems, and by engineering the project so that spare plant, test facilities, etc., are shared.

It is a desirable feature of future development that all broadband channels should be interchangeable between cable and radio and usable for telephony or television.

Large numbers of circuits will be required on main trunk routes to cater for the anticipated growth in traffic when subscribers are enabled to dial long-distance calls, a facility which it is planned to introduce in the United Kingdom, beginning probably with calls from Bristol to London and certain other centres in 1959. Long-distance subscriber dialling should extend fairly rapidly to routes between other large cities.

In order to make nation-wide dialling by subscribers practicable it is necessary to set up a national numbering scheme; this will enable connection with the wanted telephone to be made by dialling a code to reach the area and exchange of destination, followed by the subscriber's number. The code should be independent of the point of origin of the call. A system planned with this objective must be capable, in the first place, of being grafted on to the telephone system as it stands to-day, without drastic rearrangement of plant. Because of this it is impossible to bring all local exchanges, large and small, in advance into what are known as linked numbering schemes, as in Switzerland and North America. In the second place, the national numbering system must be capable of meeting the full requirements of telephone development for many years to come. This has made the Post Office anticipate a system of 20 million subscribers connected to 8 000 exchanges.

Charging for trunk calls is now based on a ticket made out by the operator in respect of each call. This practice cannot be maintained if the full economic advantages of subscriber dialling are to be realized. Automatic message accounting equipment has been developed and is being used in North America, where long-distance calls extend to much over a thousand miles, and where various other factors differ considerably from those applicable in the United Kingdom. Alternatively, where maximum distances are much shorter, charges for long-distance calls could be derived from timed operation of the subscriber's meter with bulk billing. Electronic register translators would be suitable for the dual purpose of routing the long-distance call and of determining the meter pulsing interval appropriate to the distance.

Electronic techniques are likely to revolutionize the art of telephone switching within the foreseeable future. In America, the Bell Telephone Laboratories have announced that they will have a fully-electronic exchange in public service by 1958. Many telecommunication laboratories are pressing on with the development of systems which will render the present mechanical equipments—cross-bar as well as Strowger—obsolescent, although mechanical systems with electronic control may be used as an interim measure. So far as is known, fully electronic systems are beginning to fall into two broad types. The first uses gas diodes or some other device for interconnecting the speech circuits in a space multiplex; the second, described by T. H.

Flowers and his collaborators, uses time-division multiplex for interconnecting the speech circuits and control.

Development of electronic exchanges is at an interesting stage. The philosophy of the switching has been worked out in terms of broad functional designs. The speed with which our ideas can be realized in the form of a cheap and compact exchange depends on the production of apparatus for performing the various functions. In some cases there are alternative methods for doing what is required; for example, cathode-ray tubes with thousands of tiny capacitors deposited on the screen, assemblies of cheap mass-produced ferrite cores, and delay lines are three different forms of electronic memory for storing large amounts of information. Their future relative popularity will depend on how improvements take place.

The production cost of an electronic exchange is likely to be less than that of the corresponding mechanical equipment. Smaller and cheaper buildings will suffice to house the equipment, and incidental savings in capital expenditure on local cables may be possible if the network can be adapted to the system. Maintenance costs should be appreciably less.

World Telecommunication

The concept of a transatlantic telephone cable, having submerged repeaters at intervals to increase its traffic capacity, was first described to The Institution by Dr. Buckley in the 1942 Kelvin Lecture. Sir Stanley Angwin, in his Presidential Address the following autumn, took up the theme of transatlantic communication, and the first telephone cable system was completed last August. A general description of the design objectives for the system was given in a paper read to The Institution in November, 1954, and it is not my purpose to review the results obtained on the system this evening. That will be done in January, 1957, when papers by British and American authors, describing the cable and repeaters, will be read at a joint meeting between The Institution, the American Institute of Electrical Engineers and the Engineering Institute of Canada. The meeting will be conducted, appropriately, over the cable.

For the benefit of those members who have not followed recent developments in submarine cable technology very closely, it is well to recall that it is only the advent of the submerged repeater that has made the long submarine telephone cable possible. The attenuation of the high-frequency signals required to transmit one telephone channel, let alone a number of channels over a single conductor, is very rapid as the signals pass along the cable. In the Newfoundland–Scotland section of the new transatlantic cable the loss at the highest frequency transmitted, 164 kc/s, is about 1·6 dB per nautical mile—at this rate of attenuation the output of a 100 MW generating set could not be detected by the most sensitive galvanometer more than 150 miles away. The loss is made up by the insertion of a repeater with a gain of 65 dB—a power gain of over a million times—every 38 miles. The combination of cable and repeaters gives a system free from loss, but which depends on the continuous operation of a great deal of electronic equipment at the bottom of the ocean.

Completion of the project marks the opening of a new era in the growth of world communication. It would be appropriate, therefore, to review the usefulness to this end of long deep-water cables equipped with repeaters. The development is revolutionary in its possibilities because the capacity of this kind of cable is likely to exceed that required to replace all the existing telephone and telegraph facilities on its route, and the intention of the American Telephone and Telegraph Company to lay a replica of the transatlantic system from the United States to Hawaii, half-way across the Pacific, in 1957 means that another big step towards the development of a global network of the new kind has already been taken.

The new development in world communication must be of very considerable interest to British engineers for a variety of reasons. In the first place the United Kingdom has not only a unique tradition in the manufacture of submarine cable, but also a manufacturing capacity which is unmatched in any other country and has been greatly increased and modernized in the last few years. Next, British research on thermionic valves, and British circuit techniques are likely to exert a major influence on submerged repeater practice during the next decade. Lastly, Cable and Wireless Limited owns 150 000 miles of submarine telegraph cable, more than half the world's total mileage. Communication facilities provided by telegraph cables are limited and expensive compared with what is inherent in a modern cable with repeaters, but the old and new facilities must exist side by side, or in combination, for many years. This will create problems of use in which the Commonwealth Governments will be concerned.

In order to study the potentialities of long repeatered submarine-cable systems it is best to start from the traffic requirements, expressed in terms of the maximum frequency that must be transmitted over the cable. I have chosen 3 kc/s as the requirement for each telephone channel as a compromise between the 4 kc/s internationally standardized for circuits on land and any more economical arrangement that may be possible later using unconventional band-compression techniques which are being intensively studied. Eighteen telegraph, or telex, channels may be substituted for any one of the speech channels by the use of appropriate terminal equipment.

The relative merits of double- and single-cable systems have been much debated between British and American engineers. The use of separate cables for the two directions of transmission makes for simplicity of repeater design, and the repeaters can be accommodated in flexible housings not much larger than the cable in diameter. A single cable, transmitting in both directions, is more adaptable and, if the number of circuits required is within its capacity, will always provide the cheaper system; it is clearly advantageous when a comparatively small number of circuits is required. With a single cable, 'go' and 'return' speech channels are separated on a frequency basis and the filter elements necessary to do this in each repeater add to the complexity of the circuit and very considerably to the space required. The more commodious rigid repeater housings which are comparatively difficult to handle are therefore regarded as essential with both-way cables. Where a single both-way cable would not meet the traffic requirements, the factors just mentioned, and the frequency bandwidth which is lost in a both-way cable between the two directions of transmission, would be reasons for not using two both-way cables in preference to two one-way cables. On the other hand, there are obvious advantages in providing two independent systems where adequate alternative routing is not available.

Whichever kind of system is chosen, it is possible to work out an optimum design. As the diameter, and therefore the cost, of the cable is decreased, its attenuation increases and more repeaters are required for the transmission of the frequency bandwidth to carry the traffic.* At the present stage it is probably easier to predict the electrical performance of a combination of cable and repeaters than the future price of either. In particular, the cost of submarine cable is very largely dependent on the price of the basic raw materials, copper, polyethylene and steel. The proper cost of repeaters is still difficult to determine; but it is safe to

* For standard coaxial cable with polyethylene dielectric the attenuation varies approximately as $\sqrt{f}/d$, where f is the maximum frequency transmitted and d the diameter of the core. The working gain obtained from each repeater is limited because of restrictions on the minimum permissible input level and maximum permissible output level for satisfactory performance.

conclude that prices will remain high because of continuing high research and development charges and because of the care required in the control of manufacture and in the selection and testing of materials and components.

Fig. 2 shows the minimum capital cost of single- and double-cable schemes designed to meet various traffic requirements on a route 2000 miles long. The traffic capacity is shown in terms of the frequency bandwidth available for each direction of transmission. The calculations on which these curves are based take into account the costs of the heavily armoured cable used at the ends of the system, where the cables are in comparatively shallow water, and also of the terminal installations. It must be emphasized, however, that the costs shown are not those of actual systems; the curves are only intended to indicate the variation and order of magnitude of costs.

Annual charges are of greater importance than first cost, and to obtain these the cost of maintenance, including occasional repairs and replacement of repeaters, has to be added to the interest and depreciation charges on the various parts of the system. Fig. 3 shows the estimated annual cost of a telephone circuit in 2000-mile cable systems of different capacity. The assumption has been made that the system will have a life of 20 years, after which the total investment is written off; also that one repeater in every ten will have to be replaced in addition to the normal incidence of cable faults.

For the design data which have made Figs. 2 and 3 possible, I am indebted to R. J. Halsey, who has carried much of the British technical responsibility in connection with the transatlantic telephone cable.

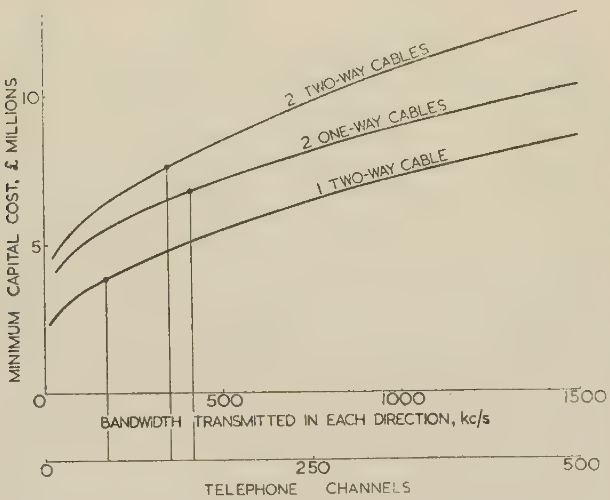


Fig. 2.—Capital costs of cable systems: 2000 n.m.
The three ordinates indicate limits of capacity set by present construction technique.

It will be seen that the cost per circuit falls rapidly as the capacity of the system is increased. A tenfold increase, from 10 to 100 circuits, decreases the cost per circuit by a factor of about 6; another tenfold increase to 1000 circuits would decrease the cost by a further factor of 4, or 24 times in all. It is clear, then, that the way to comparatively cheap circuits lies in using cables having a large traffic capacity. Economically, provision for all facilities, telephony, telegraphy, telex—and ultimately television—should be combined in one cable.

There are practical limitations to the kind of system we can build at present. These are set by:

(i) The maximum voltage which can be applied to the system to energize the repeaters.

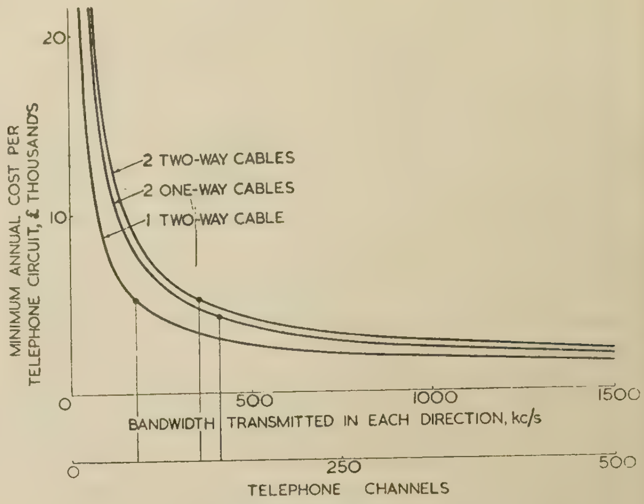


Fig. 3.—Annual costs of cable systems: 2000 n.m.
The three ordinates indicate limits of capacity set by present construction technique.

(ii) The minimum spacing at which it is practicable to insert repeaters in the cable.

(iii) The maximum size of coaxial cable it is practicable to lay and recover for repairs.

The second and third of these limitations depend on the depth of water in which the cable is laid. All three of them present problems—in high-voltage, mechanical or marine engineering, and it will be of interest to examine the three in turn.

The 51 repeaters in each of the new transatlantic cables are energized in series by a direct current fed along the centre conductor in common with the high-frequency speech currents. This warms the valve heaters and provides the anode voltages. A total driving voltage of about 4000 is required, the centre conductor and all the connected apparatus in the repeaters being 2000 volts positive to the sea at one end of the cable and equally negative at the other. Repeaters could certainly be adapted for 3000-volt working, and there is no real reason why 6000 volts should not be possible. With valves having practicable heater designs and using low anode voltages this would permit of repeater spacings of less than 15 n.m. on a 2000-mile cable and would cover all likely requirements, apart from television, for a long time to come.

During the last two summers, repeaters in flexible housings have been laid without incident at intervals of 38 n.m. across the Atlantic, one of them at least in a gale which would have made any cable-laying operation difficult. This kind of housing is capable of passing through the normal ship's cable gear during a continuous process of cable laying and adds very little to the hazards of cable operations. It is, however, not only inadequate for both-way repeaters but provides no room for duplication of circuit-elements or components in any repeater. Use of a duplicate amplifier, which is recent British practice, reduces the risk of failure and is particularly desirable if advantage is to be taken of new components or valves with only a short life history. It becomes therefore necessary to plan the use of rigid repeater housings generally. In shallow waters, comparatively large rigid housings have been successfully laid at intervals of 16 n.m. The methods adopted required the manual help of practically all the deck crew of the cable ship and are not suitable for operations extending over many days or in bad weather. Laying machines differing entirely from the proved and conventional drum have been suggested. Models of some have been built, and at least one has been installed on a cable ship. But there is no real experience yet as to how reliable they will be working con-

inuously under adverse conditions at sea. Experience with rigid repeaters in deep water is very meagre. Figs. 4 and 5 illustrate the present handling of flexible and rigid repeater housings.

Fig. 4.—Flexible repeater passing through cable-laying gear.

Fig. 5.—Rigid repeater being taken past cable-laying gear.

Quite apart from any initial laying difficulties, however, replacement of any kind of repeater in mid-ocean may mean adding five or more miles to the cable length and a supplementary repeater. Until more experience has been gained, therefore, 20 to 30 n.m. should be regarded as the minimum spacing of repeaters in deep water.

Much has been learned about the laying and recovery of submarine cables during the past century, but the margin between success and failure often remains very small. The 1956 transatlantic cables are of the coaxial type required for telephony. They have a core diameter of 0·62 in and weigh about 3 tons per mile in air. They were laid without difficulty at 2400 fathoms. The same type of cable will be used next year in the Pacific, where the depth will approach 3000 fathoms (3 miles). With fair weather conditions the recovery tension from this depth is about 70% of the breaking strength of the cable and may approach the maximum permissible load on existing cable gear. Movement of the ship in bad weather will rapidly increase these forces. Cable has to be recoverable if repeaters are to be replaced, and an operation started in fair weather may have to be completed in foul. Given suitable gear, heavier cable could be used in deep water, but, with existing gear, cable with 0·62 in core should be regarded as the largest for planning purposes.

It is well known that in laying conventional cable, and to much greater extent in its recovery, stored turns in the armouring wires constitute a major hazard which is certainly not reduced when repeaters are present. Research has been undertaken on a form of structure with high-tensile-steel core which is torsionally balanced. The cable has other attractions in respect of weight and cost.

The limits set by 20-mile repeater spacing and 0·62in diameter cable are shown in Figs. 2 and 3. Because of the separate limits, achievement of maximum traffic capacity may require departure from the cheapest design, but in most cases this does not greatly increase the cost.

In the present state of the art, transistor amplifiers would compare unfavourably with valve amplifiers in respect of power-handling capacity and noise level and would be only marginally suitable in other respects. The art is, however, a progressive one, and when, as is likely, transistors of improved performance and proven reliability become available they will increase the prospects of building long cables with sufficient frequency bandwidth for the transmission of television.

The impetus given to the growth of communication between Western Europe and North America by the 1956 cable may well call for the laying of a second cable across the North Atlantic at no distant date of the greatest capacity technically possible. Eighty circuits in a single cable, or 200 in a twin-cable system, would be an objective only just outside the present limits of repeater spacing and cable diameter. If reasonably loaded with traffic at present call rates, either system would be profitable. The advances that are taking place will make still larger systems possible in a few years' time.

On other world routes the prospects are different. From a comparison of populations, circuit requirements to Australia and New Zealand are probably not more than between 5 and 10% of those to North America. Twenty-four circuits provided over a single-cable system 12000 miles long would require about 550 repeaters with an even higher standard of fault immunity than that postulated earlier. The annual charges on each circuit would be reduced if circuits were provided for part of the way in larger-capacity cables. With costs as high as they would be, cable clearly cannot be competitive on cost with radio for the provision of small groups of circuits for very long distances. Except for the use of relay stations on some radio routes, the cost of terminal transmitters and receivers is about the same for all, and this has set a pattern of uniform charges to most destinations beyond Europe.

Much has been done to improve the reliability of long-distance radio circuits, and techniques have been developed whereby more channels can be accommodated in the high-frequency band (3–30 Mc/s), which has generally been used for such circuits. The limit to the number of long-distance circuits that can be used in this band makes it certain, however, that it alone cannot

satisfy the growing need for world communication. Neither has it been possible to achieve complete reliability. Despite the improvements, it does seem clear that the propagation of high-frequency radio waves is such that continuous communication cannot be realized over such difficult routes as the North Atlantic, or to the Antipodes.

Within the past few years considerable development has also been carried out on radio propagation using 'scatter' techniques, whereby weak but consistent signals can be received well beyond the horizon. Experience suggests that links relying on forward scatter from the ionosphere or the troposphere may be more reliable than those provided by more conventional means. Unfortunately, very high transmitter powers have to be used, and highly directive, and therefore very extensive, aerial arrays are required at both ends.

Tropospheric scattering of ultra-high frequencies can be used to provide a broad-band path suitable for the transmission of television. The range is at present limited to around 200 miles, but it seems clear that successions of 'scatter' links in tandem may contribute to the growth of international communication. Apart from the possible transmission of television, however, such methods do not appear competitive with the more conventional high-frequency point-to-point radio or cable on long, world routes.

The challenge to the communication engineer is to move towards the provision of cable facilities on the main world trunk routes on a scale which has hitherto been regarded appropriate only on land. He has increasing confidence in the use of electronic equipment in quantity at inaccessible points under the ocean. What is required is a means for laying in deep water the rigid housings for this equipment without all the hazards which attend such operations at present. Development of some form of mechanical gear for doing this as part of a continuous cable-laying operation is the immediate task.

A repeatered cable from the United Kingdom to Gibraltar would appear to be a useful first step in building up communication facilities on routes other than the North Atlantic. Traffic to the Iberian Peninsula alone might not justify the investment, but the cable would have potentialities for further extension—to West Africa and South America via the Azores. A technical and economic study is being made.

Conclusion

I have departed from the tradition of the last few years in that I have not mentioned in the course of this Address our great need for more engineers and technologists, or the means of training them. I have instead described a few of the things that are waiting to be done by engineers—things in which I, myself, am particularly interested. If an echo of what I have said reaches some who are still on the brink of a career and kindles in them enthusiasm for electrical engineering, or even gives an added impetus to those of us who are already embarked on its practice, I shall be satisfied. The profession represented by this great Institution can do much to serve the Nation, the Commonwealth and the world at large.

DISCUSSION ON 'AN X-BAND MAGNETRON Q-MEASURING APPARATUS'*

Mr. J. R. M. Vaughan (*communicated*): Some private discussions on this paper have shown that there are in use two different definitions of the coupling of a directional coupler, and that this fact is often not realized, so that confusion may arise. The definition used by Mr. Twisleton is that of Riblet and Saad† relating the coupled power to the power in the main guide *output* arm; referring to Mr. Twisleton's Fig. 5:

$$K = 10 \log P_4/P_2$$

In this case an equipartition coupler ($P_4 = P_2$) is described as a 0 dB coupler.

The other definition is that of Montgomery‡ referring the coupling to the *input* power:

$$K = 10 \log P_4/P_1$$

The equipartition coupler is now called a 3 dB coupler (or -3 dB by some engineers).

An actual example of confusion occurring may be seen in a paper by Mr. E. M. Wells,§ where Riblet's definition is used on a graph and Montgomery's (by implication) in the text.

If Riblet's definition is used in connection with a coupler such as Tomiyasu and Cohn's 'transvar',|| the possible values range from $+\infty$ to $-\infty$, so that a sign convention is also required. With Montgomery's definition the possible values do not pass through zero, so no confusion arises from lack of a sign convention.

If the definitions are extended to cover such devices as magic tees and hybrid rings, the question which is the main and which the secondary guide becomes obscure, so that application of Riblet's definition becomes still more confusing.

A rough poll of some colleagues showed general support for Montgomery's definition, few realizing that any alternative existed. I suggest, therefore, that Montgomery's definition should be generally adopted, and Riblet's allowed to lapse.

Mr. J. R. G. Twisleton (*in reply*): I am grateful to Mr. Vaughan for pointing out an inadvertent error in the definition of directivity in Fig. 5 of the paper, and regret that this may have added to the confusion he mentions. The directivity should be defined as $P_3/P_4 = K_2/K_1$. I agree that Montgomery's definition of coupling is more logical than Riblet's, and it also appears to be better known. Perhaps the sentence preceding Fig. 5 should be reworded 'The directivity and coupling factors may be defined as the ratio of powers obtained with all the arms matched, as shown in Fig. 5', since both Montgomery's and Riblet's definitions of coupling are used in the paper.

Referring to Fig. 5, K_1 is Montgomery's definition of coupling, whilst P_4/P_2 is Riblet's definition. The former was appropriate in the analysis of Section 5.1, whilst the latter was introduced to enable the performance of the directional couplers to be compared with that quoted by Ribler in his paper. Both definitions are used in subsequent formulae, e.g. eqns. (10) and (11).

Taking account of the error in the definition of directivity, eqns. (11) and (13) now become

$$R = \sqrt{\left(\frac{D}{1 + C}\right)} \quad . \quad . \quad . \quad . \quad . \quad (A)$$

$$\frac{\Delta V_4}{V_4} = \pm r_2 \sqrt{\left[\frac{D}{(1 + C)^3}\right]} \quad . \quad . \quad . \quad . \quad (B)$$

* TWISLETON, J. R. G.: Paper No. 2037 R, May, 1956 (see 103 B, p. 339).
† *Proceedings of the Institute of Radio Engineers*, 1948, 36, p. 61.
‡ 'Techniques of Microwave Measurement', M.I.T. Radiation Laboratory Series, 11, p. 859.
§ *Marconi Review*, 1954, 3rd quarter, p. 86.
|| *Proceedings of the Institute of Radio Engineers*, 1953, 41, p. 922.

Paper No. 2257
Jan. 1957

MEASUREMENT AND CONTROL SECTION: CHAIRMAN'S ADDRESS

By DENIS TAYLOR, M.Sc., Ph.D., Member.

'THE MEASUREMENT OF RADIOACTIVITY'

(*Address delivered 9th October*, 1956.)

My subject should need little in the way of justification. I shall be talking about 'Measurement', which is one of the key words in the title of our Section, and I may even mention 'Control'. The study of radioactivity and its measurement was at one time the province of the physicist, and perhaps the chemist also, but the industrial development of atomic energy has introduced many changes, and now it is equally the province of the electrical engineer. It seemed quite appropriate, therefore, bearing in mind my particular responsibilities in the Atomic Energy Authority, that I should address you on 'The Measurement of Radioactivity'.

Basis of Measurement

A radioactive substance, as is well known, disintegrates with the emission of nuclear particles (α-particles or β-particles) and/or photons (usually γ-rays), and the disintegration rate, which may be determined by measuring the rate of particle (or photon) emission, is decided by the amount of radioactive material present in the sample under investigation and by its half-life. Hence, measurement of the emission rate provides a convenient method for the assay of such materials.

The rate of particle (or photon) emission is determined by placing a radiation detector in close proximity to the radioactive sample. It is unusual to employ a detector which completely surrounds the sample, i.e. with 4π geometry, except in special cases, so that, in general, only a certain fraction of the particles (or photons) emitted by the radioactive source actually enters the sensitive volume of the detector. Moreover, in general, only a certain proportion of these will eventually be detected, since the detection efficiency may be less than 100%. Nevertheless, it is usually possible to ensure that the fraction of particles (or photons) which enter the sensitive volume of the detector unit and are detected remains constant from one experiment to another. Hence, it is quite feasible to calibrate a measuring instrument of this kind to read directly the amount of radioactive material present in a given sample. This is, in fact, the method which is normally adopted.

Instruments for the measurement of radioactivity present design problems of the same nature as occur with the more conventional measuring instruments. An example will make this clear, and the example I have chosen is an instrument for the continuous monitoring of α-active material in solution in the presence of γ- and β-active matter also in solution. It comprises a scintillation counter, a liquid-sampling device and a system of standardizing against a standard radioactive source.[1] The scintillation counter comprises a screen of zinc sulphide as the phosphor and a photo-multiplier tube, both mounted in light-tight compartments. The light-tight covering, or window, over the screen has to be sufficiently thin to allow the passage of the α-particles to the screen, which means a window thickness of about 1 mg/cm.[2] The light from the phosphor scintillations is transmitted through a polished Perspex rod to the photo-sensitive layer inside the end-face of the photo-multiplier tube. This allows the latter to be mounted in an accessible position away from the active liquor.

The system is illustrated in Fig. 1, from which it will be apparent that the liquid is brought into close proximity to the detector, by means of a revolving drum which dips into the active liquid. This is a convenient method of ensuring exactly reproducible geometry, a necessary condition for proportionality between the measured counting rate and the amount of α-active material present in the sample. The rotating disc between the drum and the detector carries a standard radioactive source over one half, and the other half is provided with a window through which the α-particles from the liquid can pass and be detected. With this arrangement, first the liquid and then the standard source is presented to the detector, the whole cycle repeating itself with every revolution of the disc. The trains of pulses are amplified and then pass on to a pulse-amplitude discriminator which permits only those pulses above a definite threshold value to be passed to the counting apparatus. The counting system is a two-channel counting-rate meter which is switched in synchronism with the disc rotation and gives two separate meter indications—one of the activity of the process solution and the other of that of the standard. The indications are recorded on a self-balancing potentiometric recorder to give a continuous indication, not only of the two activities, but also of their ratio, i.e. the unknown in terms of the standard. This is useful, not only because of the automatic calibration feature, but because the separate recording of the measured activity of the standard source allows any faults in the apparatus to be quickly revealed.

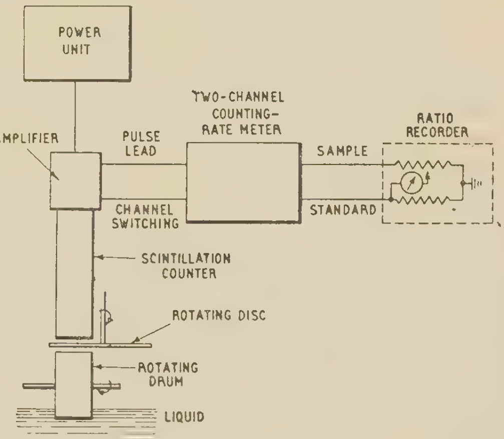


Fig. 1.—Instrument for measuring α-activity in solution.

Ultimate Sensitivity

The ultimate sensitivity of radioactive assay can be very great. An example will make this clear. Let us suppose that we wish to assay radio-phosphorus (P^{32}), which emits β-particles. A Geiger counter can be used as the radiation detector in this case,

Dr. Taylor is at the U.K.A.E.A. Atomic Energy Research Establishment, Harwell.

and such counters detect β-particles with an efficiency of very nearly 100%. The background counting rate of a detector of this type of the usual size when shielded by 2 in of lead is of the order of 12 counts per minute. If we assume that we can detect a doubling of this counting rate with the source in position beneath the detector, and the geometrical arrangement is such that one-fifth of the particles pass into the detector, then the β-particle emission rate will be $12/60 \times 5 = 1$ per sec. If, moreover, the disintegration scheme is such that one β-particle is emitted per disintegration, one per second is also the disintegration rate. This is a very small quantity, as can be seen by using Avagadro's hypothesis and the law of radioactive decay.* In this case it is only $9{\cdot}41 \times 10^{-17}$ gramme (corresponding to $2{\cdot}7 \times 10^{-11}$ curie for P^{32}).

In practice the ultimate sensitivity is determined very largely by the background counting rate. Reduction of this rate, which allows higher sensitivity, may be achieved by using special counters which have been carefully constructed to exclude materials containing traces of radioactive materials and using a small ring of further counters in an anti-coincidence circuit[2] to remove the effect of the cosmic rays. In this way it is possible to gain a factor of about 10 times and achieve a background counting rate of only 1 count per min.

Where the experimenter has the choice it is usually better to measure an α-activity rather than a β- or γ-activity if the highest sensitivity is the aim. This is because counters sensitive to α-particles can be made which have a background counting rate of only 2 counts per hour.† In these counters, the residual counting rate is due to small amounts of activity which occur as impurity in the materials of construction, plus any activity in any air volume counted. It is not possible to use the anti-coincidence method to achieve further improvement in this case, since none of the residual counting rate is due to cosmic radiation. However, the detectors usually employed for α-activity measurements (gas proportional counters and scintillation counters) are capable, in association with suitable pulse-amplitude analysers, of selecting events in any desired energy range. It is therefore possible to arrange the channel setting to embrace the energy peak of the required emissions, and then only the background

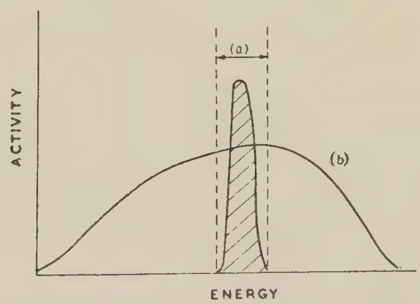


Fig. 2.—Use of single-channel pulse-amplitude analyser to reduce background counting rate: spectrum of energies.

(*a*) Energy emitted by substance to be assayed.
(*b*) Energy contributed by background radiation.

corresponding to this energy range will be counted. This is illustrated in Fig. 2. In general, an improvement of about 10 times is possible by this means.[3]

* Using Avagadro's hypothesis and the law of radioactive decay, it works out that the mass of material in grammes is

$$2{\cdot}38 \times 10^{-24} AT(dN/dt)$$

where A is the atomic weight of the material, T is the half-life of the material and dN/dt is the disintegration rate.

† In practice the natural activity obtainable can be as low as that corresponding to a mean counting rate of one α-particle per hour for 10 cm² of exposed surface of the walls of the counter.

Measurement of Body Radioactivity

A problem which has come to the fore in recent years is the measurement of the radioactivity of the human body. The efforts expended have been mainly directed towards measuring the natural radioactivity of the body and to devising techniques for the early detection of its increase. There are several reasons for this, the chief one being the necessity of ascertaining at an early stage whether a person working with the radioactive substances has been contaminated by them.

The extremely low value of the maximum permissible total body burden of many nucleides laid down by the International Committee on Radiological Protection (I.C.R.P.) as a safe maximum quantity (1×10^{-7} curie for radium and 4×10^{-8} curie for plutonium) presents a measurement problem of real difficulty, especially as it is really necessary to make measurements below this level. Furthermore, it is also necessary to make measurements of subjects who have not been exposed to radioactive contaminants and whose body burden largely results from the natural potassium content. This latter involves measurement down to 1×10^{-8} curie or less.

As the subject involves small numbers it is perhaps appropriate to note that this quantity, 1×10^{-8} curie, is large compared with that mentioned in connection with the assay of radiophosphorus. However, in the measurement of body radioactivity, we are concerned with determining the body burden by measuring the external radiation (almost always γ-radiation, but see below), which has to pass through the body, where it is attenuated, before it can be detected. This, together with the large space occupied by the 'source', makes the measurement difficult.

It is possible to determine the body burden by excretion studies, but this method depends upon having data available on what fractions are excreted. The method which has been used more widely, namely an external radiation measurement, necessitates making assumptions about the location of the contaminant and the absorption of radiation within the body.

One of the methods in this category is that using an array of high-pressure ionization chambers as the detector system in association with an electrometer valve or vibrating-reed electrometer. Systems of this sort are in use in the Radiophysics Laboratory, Stockholm (design due to Sievert[4]), in the Finsen Laboratory, Copenhagen (design due to Taylor and his collaborators[5,6] at the Atomic Energy Research Establishment, Harwell), and at Leeds University (design due to Spiers and Burch[7]).

To obtain the highest sensitivity with these ionization chambers it is necessary to eliminate the effect of any α-particles emitted from radioactive contamination in the constructional materials of the chambers. This is done by using a collector electrode of small diameter and operating the chambers at a high pressure. The ions produced by α-particle emission from the outer electrode recombine before collection, and do not therefore contribute to the measured ionization.

In the system designed by Sievert, an array of ionization chambers is used to form a 4π detector, and this is installed in an underground cavern beneath 50 metres of solid granite. There is also further local shielding provided by about 1 metre of water in tanks round the measuring chamber system. The person to be monitored is placed horizontally inside the ring of chambers and his positioning is not then critical. This is because only the most penetrating of the cosmic rays reach the measuring chambers, and absorption of the cosmic radiations by the person being monitored is negligible in this case. However, the two other installations, at Copenhagen and Leeds respectively, are at ground level. A second set of ionization chambers was used, connected in opposition to the first set to counter systematic

changes in cosmic-ray intensity. The background current was then very nearly zero.

A typical record obtained with this type of equipment is shown in Fig. 3. It will be noted that the radiation emitted by the person

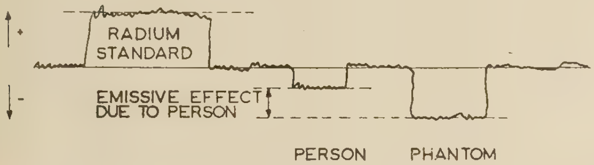


Fig. 3.—Typical results with ionization-chamber system of body monitoring.

being monitored is masked by the absorption of the body of external radiation. For this reason it is necessary to make separate determination of this absorption effect using a phantom. In both the Leeds and the Copenhagen installations this effect is reduced as far as possible by placing all the measuring chambers above the person being monitored.

All three installations have about equivalent sensitivities. Thus, with the Finsen Laboratory installation it has proved feasible to measure 2×10^{-9} curie of radium (Ra^{226}) in equilibrium with its decay products with an accuracy of $\pm 20\%$ in a 2-hour observation.

The original installation at the Finsen Laboratory was concerned with an investigation[8] of patients with internally deposited thorium. It was found that the differential (background) current increased immediately after a measurement by as much as $1{\cdot}0 \times 10^{-14}$ amp. This was found to be due to solid decay products of thoron attaching themselves to the metal lining of the patient's tunnel, the thoron having been exhaled by the patient during the measuring period. This problem was solved by fitting an extractor fan and ventilating shaft and removing the air continuously from the tunnel.

However, the most interesting measurement which has been made with this apparatus is that of the natural radioactivity of persons who have not been exposed to radioactive materials. The normal body burden in such cases is mostly due to the potassium isotope K^{40}. Since naturally occurring potassium, of which $0{\cdot}112\%$ is the radioactive isotope K^{40}, constitutes $0{\cdot}25\%$ of the lean body weight, a 70 kg man contains about 135 g of potassium, and this corresponds to a γ-activity of 1×10^{-8} curie, and results in a differential current of less than $1{\cdot}0 \times 10^{-14}$ amp for this apparatus.

It is of interest to note that, because of the effect of varying atmospheric pressure on cosmic-ray intensity, the background current from an installation of this type varies with atmospheric pressure. Some results due to one of the author's colleagues, J. Rundo,[6] using the installation at the Finsen Laboratory, Copenhagen, are shown in Fig. 4; the apparatus makes, in fact, a good barometer.

Measurements of this type have been made more recently, both at the A.E.R.E., Harwell, and in America, using a scintillation counter equipment in association with a pulse-amplitude analyser, which allows identification of the nucleides present as well as an activity determination. It also allows some improvement in sensitivity, as already explained. In America, Anderson[9] and his collaborators have used very large liquid scintillators providing 4π geometry. The apparatus, shown schematically in Fig. 5, comprises a tank holding 140 gal of ter-phenyl in toluene, and the scintillations are detected with the aid of 108 photo-multiplier tubes (2 in diameter). The equipment gives a much improved performance over that possible with the ionization-chamber systems already described. For example, the natural potassium body burden can be measured with an accuracy

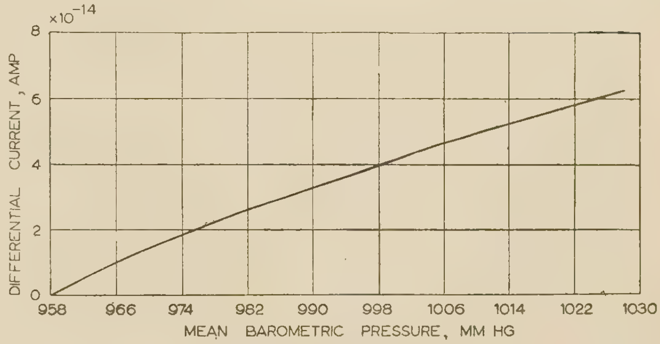


Fig. 4.—Dependence of differential current from body monitor on atmospheric pressure.

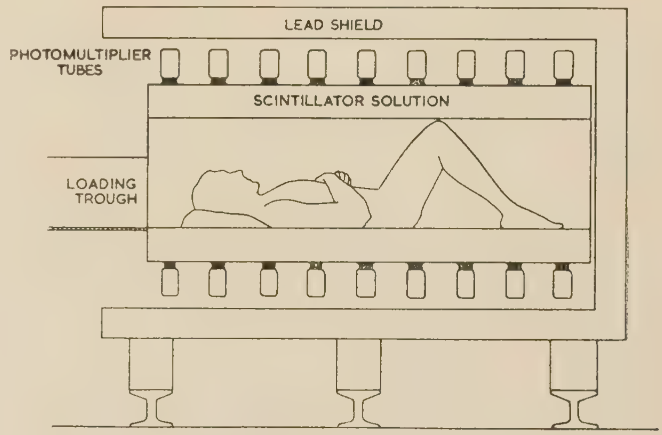


Fig. 5.—One form of body monitor using a liquid scintillator.

of $\pm 5\%$ in 100 sec of counting time, and the ultimate sensitivity is of the order of 5×10^{-10} curie.

An alternative, and it is thought preferable, system is that using sodium iodide as the scintillating material. This scintillator can be used with good sensitivity down to 10 keV, whereas the liquid scintillator probably cannot be used below about 100 keV γ-energy. This is important for some applications, as I shall mention later. On the other hand, sodium-iodide scintillating material is expensive, and it would be difficult and very expensive to produce a 4π counter with this material sufficiently large for body measurements. At the A.E.R.E. we originally used four scintillation counters each with sodium-iodide crystals, 2 in long and $1\frac{3}{4}$ in in diameter, arranged on a centre-line above the person being monitored, the spacing between successive counters being 18 in. This has now been replaced by a similar system using four photo-multiplier tubes, each employing a sodium-iodide scintillator $4\frac{1}{2}$ in in diameter and 2 in thick. With this equipment the normal body burden of K^{40} should be measurable with an accuracy of $\pm 3\%$ in a 10 min count, or $\pm 1\%$ in a 90 min count, which is nothing like as good as the liquid scintillator system but is much simpler and offers the advantage of better energy discrimination.*

One result which has been obtained recently with the A.E.R.E. apparatus is that all persons tested contained the expected amount of potassium (K^{40}), but in addition a small amount of a further isotope was detected, which has been identified as caesium (Cs^{137}). A typical set of results are shown in Fig. 6. The amount found

* This comparison on a statistical basis is not the complete story, because the stabilities of the two systems are not equivalent. With the sodium-iodide scintillation it is possible to operate on a 'plateau' and obtain good stability. This is not the case with the liquid scintillator.

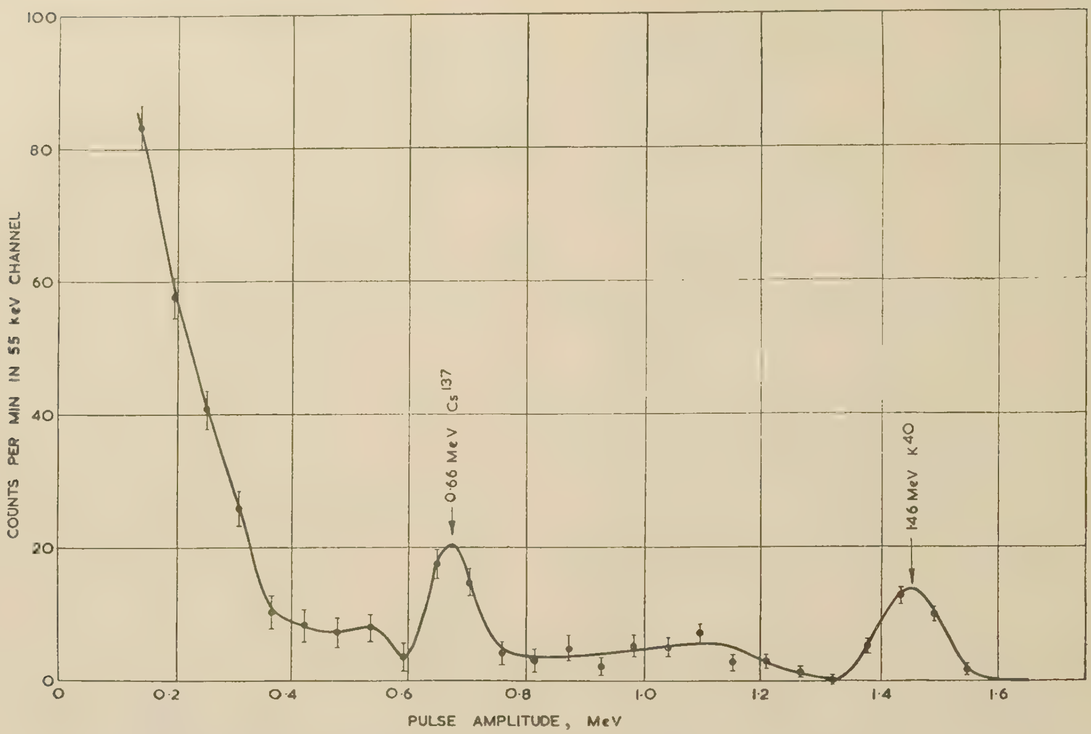


Fig. 6.—Typical results of a radioactivity measurement on a human subject.

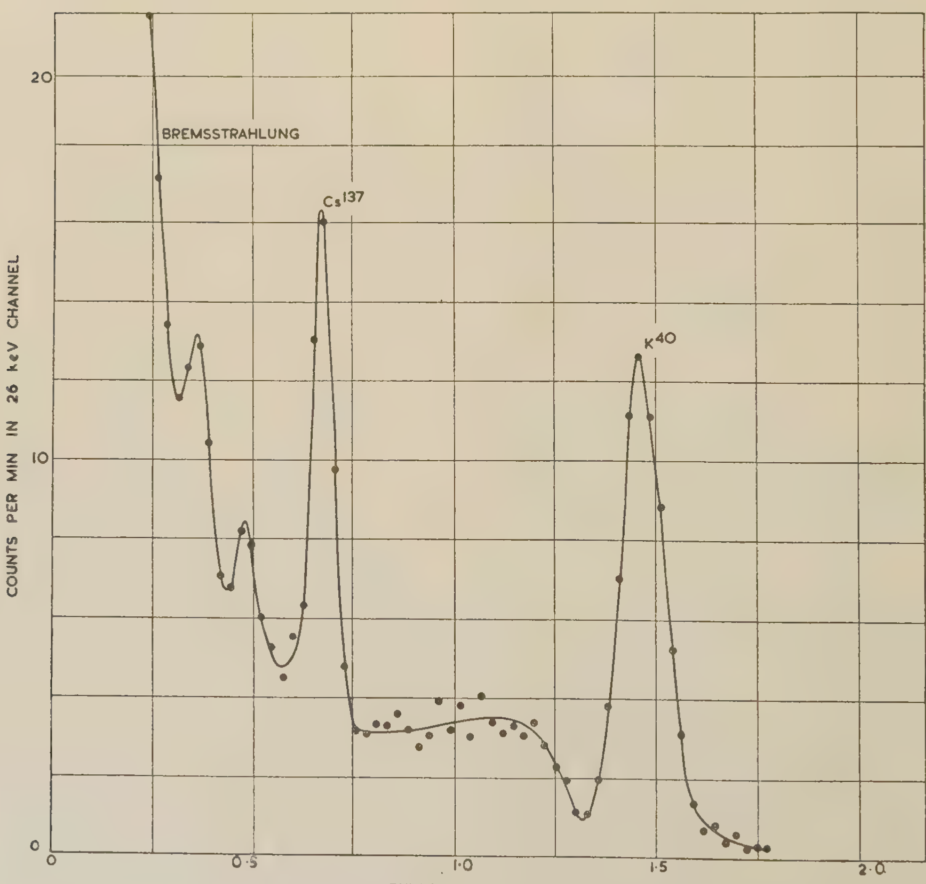


Fig. 7.—Gamma spectroscopy of a recently produced dried milk sample.

was very minute, about 4×10^{-9} curie, which is very small indeed compared with the maximum permissible body burden (90 microcuries) laid down for this isotope by the I.C.R.P. Fig. 6 brings out very clearly the use of the scintillation counter system as a tool in the atomic energy laboratory, as it not only gives the required quantitative information but also identifies the radioactive source (contaminant) by its energy spectrum. The additional peak (Cs^{137}) of Fig. 6 is almost certainly due to 'fall-out' from atomic explosions. The amount is very minute, and that it can be measured and identified at all is only because of the very high sensitivity of the spectrometer used. The 'fall-out' occurs on pasture land on crops and eventually through the cycle from grazing animals to milk into human beings. Fig. 7 shows the results of some measurements on dried milk of recent origin, and in this case, as well as the K^{40} and Cs^{137} peaks, further peaks are to be seen. Again the amounts are very minute, several orders of magnitude below the maximum permissible levels laid down by the I.C.R.P. for these materials.

As a further example, some measurements are given in Fig. 8 on carrot tops. Again the peak due to naturally occurring

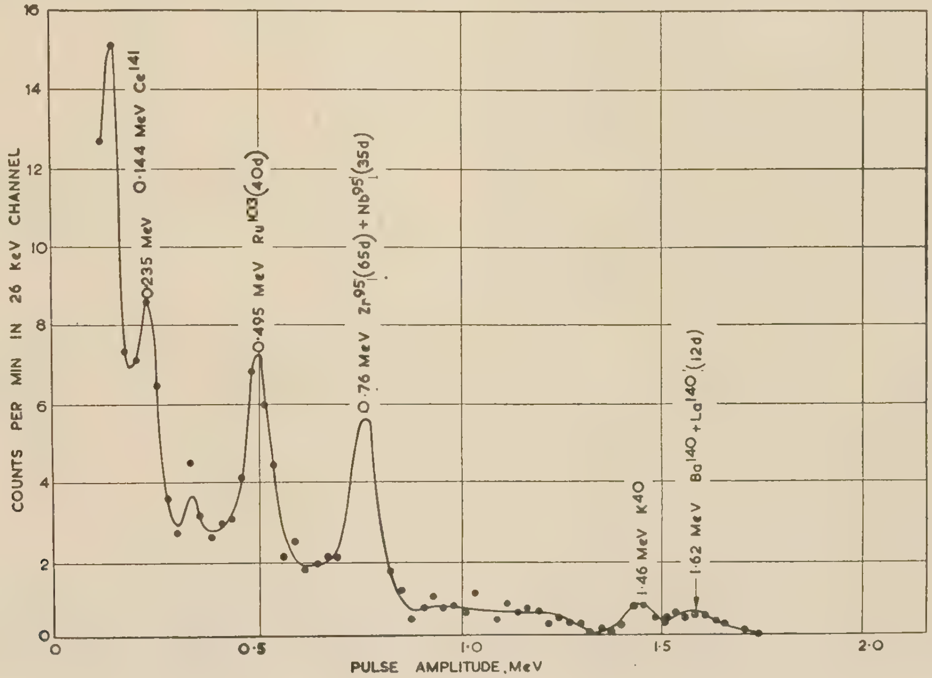


Fig. 8.—Gamma spectroscopy of a sample of carrot tops.

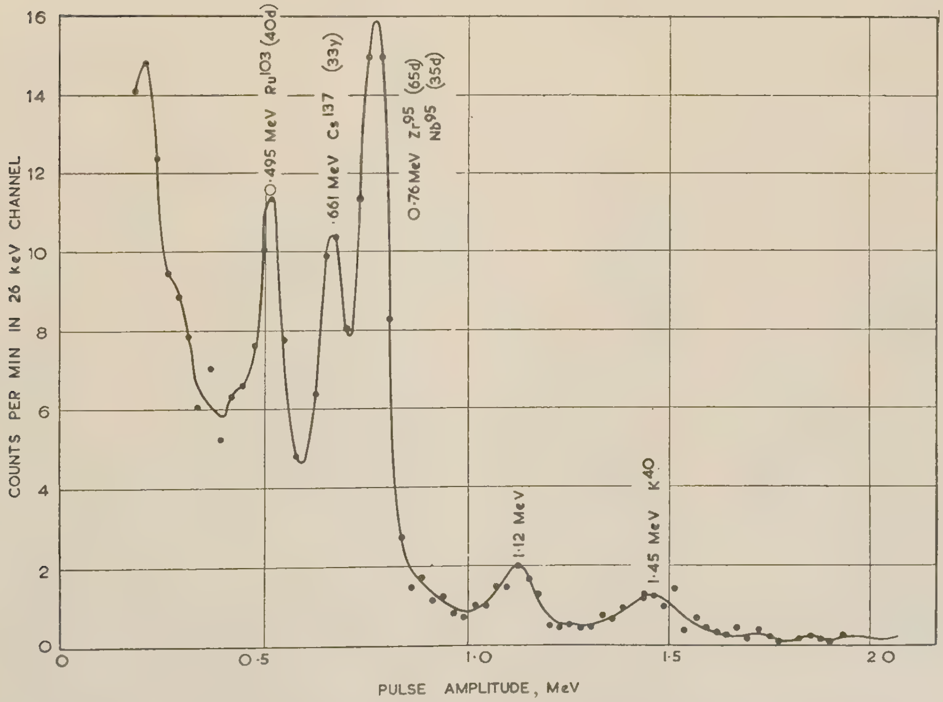


Fig. 9.—Gamma spectroscopy of domestic vacuum-cleaner dust 30 days after collection.

potassium (K^{40}) is present, but because the 'fall-out' is of very recent origin in this case, the remainder of the spectrum is mostly due to the shorter-lived fission products, and peaks will be noted due to Ce^{141}, I^{131}, Ru^{103}, Zr^{95}, Nb^{95}, Ba^{140} and La^{140}.

Although the amounts are very minute, many orders of magnitude below the maximum permissible limit, it is even possible to measure and identify activities in domestic vacuum-cleaner dust. Fig. 9 shows such a measurement.

One other small experience may be mentioned. When the body monitor at the A.E.R.E. is unshielded, the apparatus provides a method of detecting (at a distance of a few hundred yards in this case) whether Bepo, one of our larger nuclear reactors, is in operation. Pile start-up or shut-down causes a change in the counting rate of about 100 counts per sec. This is due to the very minute amounts of gaseous and volatile radioactive products which are emitted by the pile in operation, and which pass into the sensitive volume of the detector system, particularly when the wind is in a favourable direction.

A fuller description of this A.E.R.E. equipment is being published by Owen,[10] who also gives details of the expected performance with pure β-emitters (which are detected by measuring the emitted *Bremsstrahlung*), and with Pu^{239}, which it is hoped will be detected by measuring the 17 keV X-rays which are emitted in this case.

Radiation Dosage Measurements

So far, we have been discussing radioactivity measurements where the aim is the highest sensitivity. Sometimes a reasonably high-sensitivity instrument is fairly easy to design, but the production of less sensitive instruments is more difficult. A good example of this is the quartz-fibre electrometer used in association with an ionization chamber for the measurement of γ-ray dosage.

Fig. 10 shows the usual form of this instrument for the measurement of dosage in the range 0–500 mr. To obtain a range of 0–500 r, the ionization chamber is loaded with extra capacitance, and it is convenient to use a dielectric of polystyrene foil or any other suitable material. A high-resistivity dielectric is necessary for this application (of the order of 10^{20} ohm-cm) and this must be maintained after heavy irradiation (say up to 2 000 r) over the operating temperature range (up to 50° C in our case), otherwise the 'natural' leakage of the instrument makes long-period dosage measurements difficult or impossible.

It is possible to produce improved leakage characteristics in ordinary polystyrene by subjecting it to massive doses of radiation of the order of millions of röntgens. Gamma-radiation from radioactive cobalt (Co^{60}) is used for this purpose, and Fig. 11 shows the result of measurements on polystyrene samples thus irradiated for doses of 10 to 200 Mr at dose rates up to 1 Mr/h. To obtain satisfactory results, the surface of the polystyrene must be kept at a low humidity over a dessicator, and it is probably preferable to do the irradiation in vacuo.

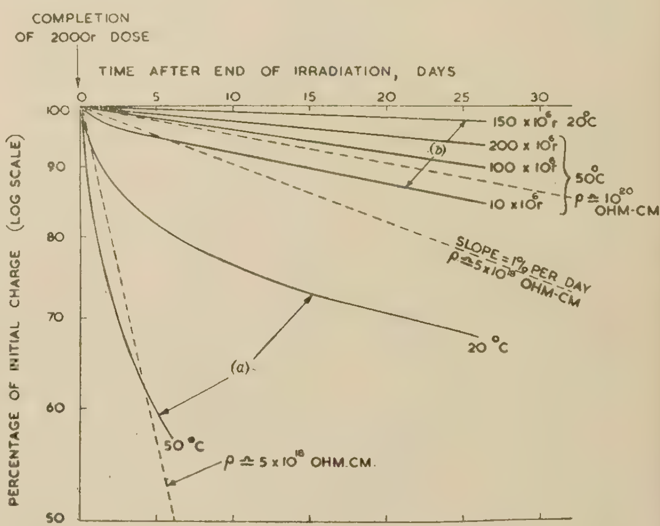


Fig. 11.—Discharge of a high-grade polystyrene capacitor (at 20° C and 50° C) due to induced conductivity after γ-irradiation (2 000 r in 2 hours):

(*a*) Without treatment.
(*b*) After treatment with megaröntgen dose.

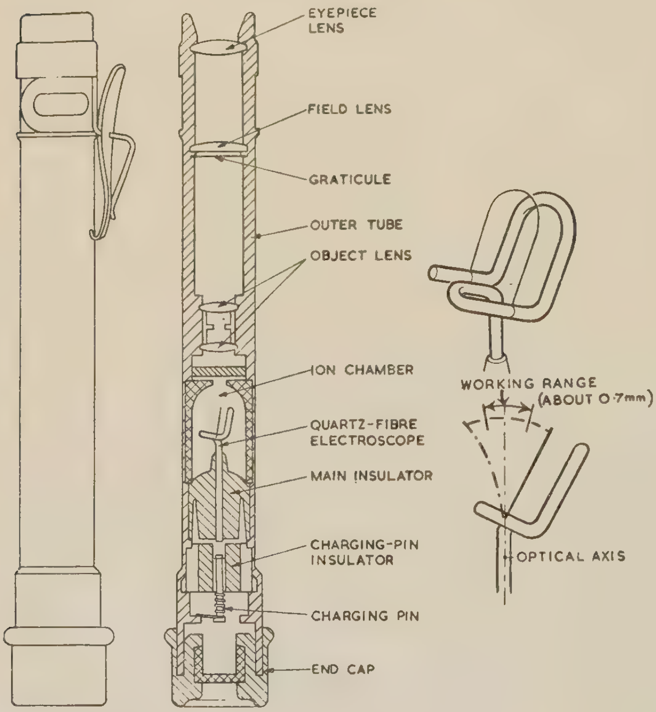


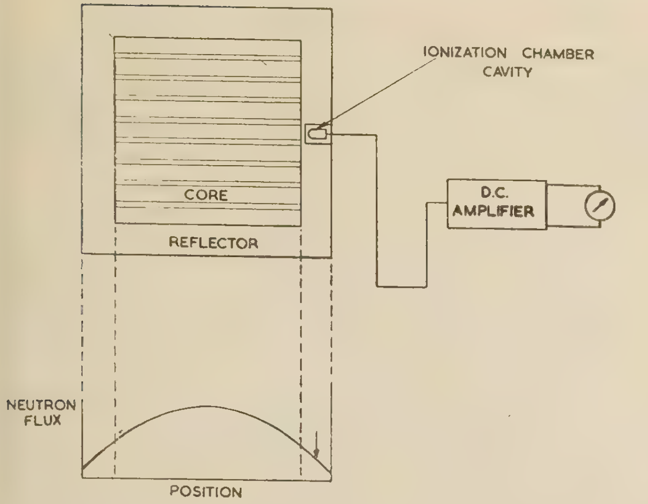
Fig. 10.—One form of quartz-fibre electrometer and ionization chamber used for measurement of γ-radiation dosage.

This method of producing high-quality insulating materials for use in high-range measuring instruments has not been used on a large scale, but is an interesting example of how the new measuring instruments required in this atomic energy era are made possible using the products (Co^{60} radiation in this case) of atomic energy.

Measurements relating to Nuclear Reactors

I would now like to discuss radiation measurements relating to reactors.

The fission rate and its rate of change in a reactor must be under control at all times from the shut-down sub-critical state to full-power operating level. When running at constant power, as has been explained by Cox and Walker in a recent paper,[11] a steady state is reached when the heat output is proportional to the fission rate. Under these conditions about 94% of the power output comes from the kinetic energy of the fission fragments and from radiations accompanying the fission process; the remaining 6% is due to β and γ-decay processes in the fission products which change relatively slowly with power level. Heat output measurements can thus be used for the measurement of fission rates only over a small range near the maximum operating



ig. 12.—Method of determining fission rate by measuring the flux escaping neutrons.

evel. A more direct measurement of fission rate is necessary or control during shut-down and start-up, and neutron-flux measuring instruments are used for this purpose. The same type of instrument is also used at full-power operation, as it is possible to obtain a much shorter response time than is possible

(ii) A pulse counter, either a fission counter, or a BF_3 proportional counter in association with a counting-rate meter for use at low levels, particularly at initial start-up.

With a pulse-counting technique, pulse-amplitude discrimination can be used to differentiate between neutrons and γ- and β-radiations, and this is often the only method which can be successfully applied at low fission-power levels. In the d.c. ionization chamber system there is no method of differentiating between neutrons and γ-radiations except the inherent neutron-gamma sensitivity.

In the paper mentioned, Abson and Wade describe an ionization chamber suitable for use over a wide range of neutron flux, and discuss the limitations of a 2-electrode chamber, using boron coating or BF_3 gas fillings, due to neutron-gamma sensitivity. The effective neutron-gamma sensitivity can be increased by the use of a 'γ-compensated' chamber, which comprises, effectively, two chambers, one sensitive to $(n + \gamma)$ radiation and the other, connected in opposition, senstive to γ-radiation only. In this ion chamber, shown in Fig. 13, the leads to the electrodes consist of concentric lines with insulators of alumina-bearing quartz, the collector-electrode lead being evacuated to minimize unwanted currents due to γ-radiation. This design is necessary because the chamber is to be used at γ-radiation levels where there would be appreciable radiation damage to polythene-insulated cables. It has also to be used under water, and then the rigid electrode-lead system is extended to a region where normal gasket materials may be used to provide a water-tight seal round flexible polythene cables. The same general technique

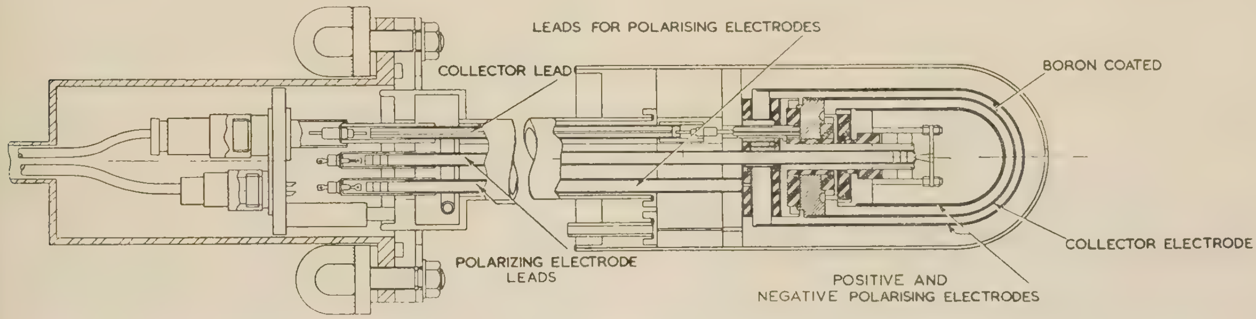


Fig. 13.—Gamma-compensated ionization chamber now being used where improved neutron/gamma discrimination is required.

n calorimetric measurements, and thus guard against sudden transients in the fission rate or fission power.

It would be very nice if the fission rate could be measured irectly, but a satisfactory method has not yet been invented. However, if it is assumed that the neutron flux pattern throughout he reactor core remains constant, a measurement of neutron ux at a particular position in or near the core can be used as an ndication of the fission rate. This is the method which has been mployed on existing reactors; it is illustrated in Fig. 12. At igh power levels the equipment can be calibrated in terms of ower output against calorimetric measurements. Separate eutron-flux measurements throughout the whole of the core may also be used for calibration purposes. This flux-scanning quipment is often required to ensure that the fuel-element oading has been adjusted to give the flux distribution necessary or efficient operation and efficient use of the fuel, quite apart om the question of calibration of control equipment.

The neutron-sensitive detectors used in present reactor-control systems have been described by Abson and Wade[12] and by Cox, Gillespie and Abson.[13] Two systems are employed:

(i) An ionization chamber in association with a d.c. amplifier for use at the higher operating levels.

might also be used in applications where the temperature at the ion chamber was too high for the use of flexible polythene cables.

It will be noted that boron 10 has been used almost exclusively in neutron-sensitive ionization chambers, but if chambers are required for use at higher flux levels it will be necessary to adopt different methods, or boron depletion troubles will limit the chamber life seriously. One possibility would be to use lithium 6, and this would offer an advantage of about 4 times. There would, however, be a corresponding reduction in the neutron sensitivity with no change in the γ-sensitivity, and so the n/γ-discrimination would be less than with the B^{10} system and this might not be acceptable. The use of a fissile material (e.g. uranium 235) probably offers the best solution to the burn-up problem. The fission cross-section and hence the depletion rate in a given flux is reduced by a factor of about 8 times over that possible with B^{10}. Hence, if longer-life detectors in higher neutron fluxes are required, there seems some possibility of providing them. However, it is probably quite reasonable to locate the main neutron detectors, as has been done at Calder Hall, outside the reactor core and to determine the neutron population by measuring the flux of escaping neutrons (see Fig. 12). If the fraction of neutrons escaping varies markedly

with time or with power level (due to changes in the flux distribution with time or power level), then it may be necessary to install flux-distribution measuring equipment to provide periodic calibration data. It seems unlikely that this equipment would ever be used as part of the control system directly. This is obviously bound up with problems of reliability and inherent safety.

The methods of neutron detection used in the control instrumentation, namely d.c. ionization chambers and pulse-type counters, may also be used for flux-distribution measurements. The difficulties associated with radiation damage and high temperatures are, of course, accentuated in measurements in the reactor core. The difficulties can be avoided by using a neutron-activation technique for the flux-distribution measurements. In this method a suitable material, usually in the form of a wire, is introduced into the reactor core and is irradiated for a specific

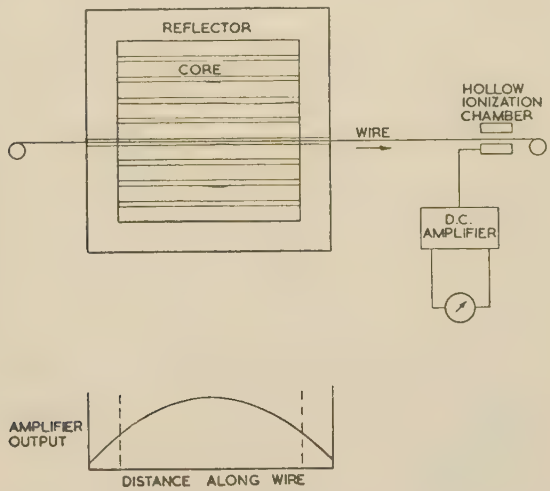


Fig. 14.—Neutron-activation technique used for flux distribution measurements throughout the core of a reactor.

time. It is then withdrawn and the induced β- or γ-activity (which is proportional to the neutron flux) is measured externally with an ionization chamber or counter. This method, which is used at Calder Hall, is illustrated in Fig. 14.

Conclusion

Only a small fraction of the radioactivity measurement problems which are engaging the attention of the U.K.A.E.A. have been mentioned in this survey, but it is hoped that sufficient has been said to indicate the scope of the measuring techniques used and the wide range of technology involved.

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Abstract No. 2313
Jan. 1957

RADIO AND TELECOMMUNICATION SECTION: CHAIRMAN'S ADDRESS

By ROBERT C. G. WILLIAMS, Ph.D., B.Sc.(Eng)., Member.

'THE ELECTRONIC AGE'

(ABSTRACT *of Address delivered* 17*th October*, 1956.)

I would first of all like to thank the Committee of the Radio ıd Telecommunication Section for my nomination, and you the embers of the Section for the honour of my election as your nairman for the ensuing year. In taking over this office I am eply conscious of the responsibilities involved in following on long line of distinguished engineers, typified by my immediate edecessor, Mr. Stanesby. Since the formation in 1919 of this, e first of the specialized Sections of The Institution, under the le of the 'Wireless Section', it has grown in professional rength and in its technical field in conformity with the enormous owth of the industry of which it is the professional engineering atform. It is my hope that in a year's time, when it becomes y turn to hand over to my successor, I will, with your help, able to look back with you on a further year of consolidation ıd growth.

It is usual on these occasions for the Chairman to take as his eme the aspects of electrical engineering with which he has en most closely associated, and in my case this is the rather oad field of production, design and technical administration uring the last twenty-five years of the electronic industry. I n proposing, therefore, to present some thoughts on its place the electrical engineering industry as a whole, some figures to ow its importance in the economy of the country, and some inters to the way in which the solid-state discoveries of the ost-war years are likely to affect the future of the industry and our profession.

The Electron as a Sub-Atomic Particle.—I am sure I cannot be one among electrical engineers in frequently pondering on the nomaly that, although electricity has so greatly changed our vilization, we know very little of what it really is. We have eveloped an accurate mathematical basis for its behaviour and n use this to design and predict the operation of large engineerg projects, but are little beyond the stage of defining it as the use which produces certain effects.

Throughout history, periods of civilization have been marked the dominant materials and techniques which have advanced ur standards of living. The 19th century saw the beginnings of e Atomic Age in which we first began to visualize the atomic ructure of matter and to utilize more fully the chemical and etallurgical properties of the natural raw materials of the world. was our leadership in these fields allied with our trading adition and our geographical location on the Atlantic seaboard at gave this country its world power. On this basis we should gically describe the 20th century as the Sub-Atomic Age in hich we have begun to understand and make use of the subomic particles, and in which every new discovery regarding the ructure of the atom is influencing the profession to which we long.

The electron happens to be the most easily freed and the most obile of the sub-atomic particles, and our profession has identied itself with it. For this reason I decided to call this address he Electronic Age' rather than what I feel is its proper title, 'The Sub-Atomic Age', since the latter would cover the technology which has become known as nucleonics.

Physics of the Solid State.—Most of the large research and development laboratories of the world to-day are occupied with the physics of the solid state, and this field of investigation is of great importance to our profession. Basic effects such as the conduction of metals, the capacitance effects in insulators, the piezo-electric phenomena in certain crystals, photo-electric and photo-conductive devices, phosphors, and the newer transistors and ferrites are all concerned with the solid state, and there seems no reason why many more interesting applications should not be ahead of us.

Solid-state work is a field in which technology has run well in advance of theory, but we are nearing the stage when theoretical analysis will enable us to synthesize the effects we require. It provides, to my mind, an example of the importance of maintaining the closest possible connections within one profession of the heavy and light divisions of our industry. Solid-state work has shown that these divisions appear to be part of an even larger profession and industry making use of sub-atomic particles and techniques of which we are now only dimly aware.

The Radio and Electronic Industry

The origin of all industries is a public demand for some amenity which has grown out of advancing technology. The electrical power industry grew primarily out of a demand for the convenience and efficiency of electric light, and in a similar way the electronic industry owes its spectacular growth to the public demand for entertainment and, we hope, instruction by sound and vision.

Size, Importance and Rate of Growth.—Although the thermionic valve dates from 1904, it was not until the First World War that production took place in any quantity and permitted the generation of continuous waves suitable for communicating speech and music. The 1920's saw the start of the public service of the British Broadcasting Company, and the enormous growth of the radio receiver market in the 1930's, together with the pioneering work of this country on television, led to a reserve of technical manpower and production facilities for mass-produced components and valves which was of vital importance to us in the Second World War.

The radio receiver industry has been unique in growing out of an already established component industry. This was due to the widespread home manufacture of simple receivers from kits of parts during the 1920's which became too complicated for the layman with the introduction of the mains-operated valve and the moving-coil loudspeaker. This enabled many new receiver manufacturers in the early 1930's to start up with very little capital, since they were able to obtain components from an industry which had previously been manufacturing direct for the public. Its present size can be measured from its employment of 200 000 persons in about 225 companies with a turnover last year of some £200 million in equipment and components, of which £34 million represented exports. Incidentally, this

Dr. Williams is with Philips Electrical Ltd.

export figure is only just below that of the United States in the same field. Census of production reports for 1950 and 1935 give the following gross outputs for the radio and telecommunication industry compared with two others with which some of us may be familiar:

	1950 Gross outputs £ millions	Ratio 1950/1935
Radio and telecommunication	137	4·9
Motor vehicles and cycles ..	669	5·1
Brewing	427	3·5

These figures show that the industry has the established capital and the ability to use this for creating the new capital from which alone further expansion can come.

The rate of growth of the industry is shown in Fig. 1.

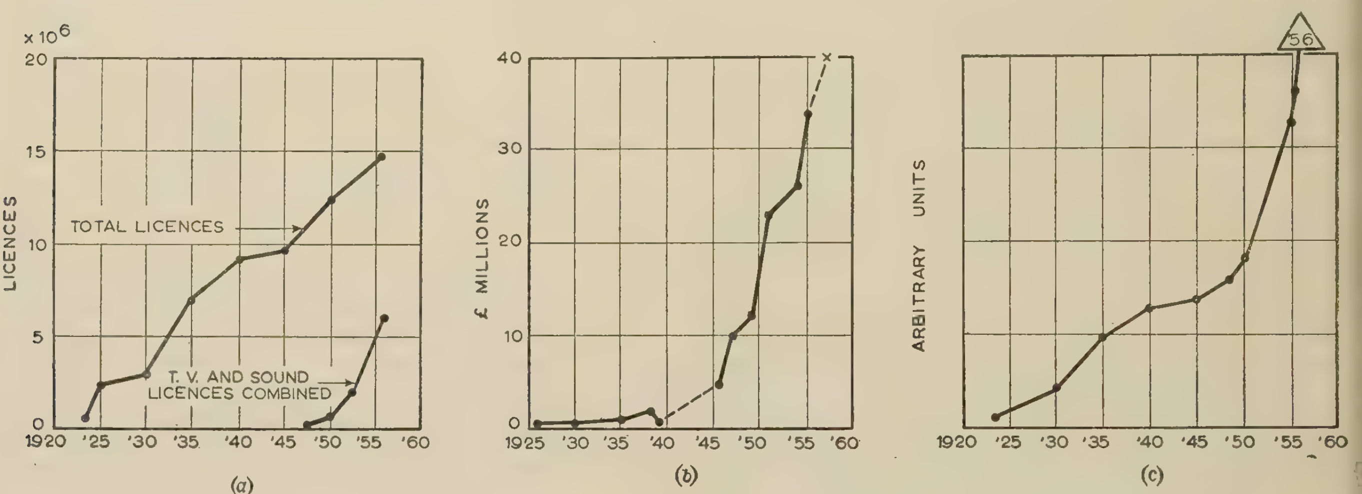

Fig. 1.—Growth of domestic licences, exports and production capacity.
(*a*) Radio and television licences.
(*b*) Exports.
×—Estimated at rate for first seven months of 1956.
(*c*) Production capacity.

export technical brain-power in the form of goods or services—in fact many people consider this to be our only real national asset. The further growth of our own industry is very dependent upon a flow of highly qualified technologists.

New developments are changing the character of the industry, and although it has grown up on consumer durables, it is predicted that in ten years' time nearly 50% of its turnover will be represented by capital goods arising from the growing demands both from home industry and the export market for sources of power and the means of controlling this power for industrial use. Whether we shall be able to take advantage of this opportunity depends largely on the quantity and quality of our professional engineers, technicians and craftsmen compared with our great competitors, the United States and Russia. A recent estimate is that the electronic industry requires about 1000 professional engineers and 4000–5000 technicians a year. As this represents nearly 30% of the present output of graduates, it is worth while reminding ourselves of the overall comparison with our chief competitors:

	Number of graduate engineers a year 1950	1954
U.S.A.	50000	20000
U.S.S.R. ..	28000	54000
U.K.	3600	3400

These figures are very disturbing but are not yet a cause for despair, since the standards are very difficult to compare and our own are as high as any in the world, but the need is obvious and calls for immediate action. The Government and this Institution have shown themselves to be fully aware of the problems involved, and the new awards in technology and the plans for extending facilities in our scientific and technical educational establishments should go far towards increasing our technological strength for bridging the gap between research and production.

Two lines of action are possible. The first is the encouragement of more students to take scientific and technical courses by means such as The Institution's film, 'The Inquiring Mind' as there is still only 34½% of our student population taking science and technology, a relatively small increase on the 26% of 1938 and 1939. The second is to take some steps to welcome rather than merely to tolerate women in these technological fields

Radio and television production figures are a useful guide to the home trade, as the domestic receiver industry accounts for a large proportion of the turnover and indicates the level of technical manpower and valve and component manufacture on which further growth is dependent. In addition, this turnover is one of the main supports of the industrial research laboratories from which come many new developments.

Our industry is no exception to the custom of companies associating together in trade organizations, and this has been of particular value in view of the very rapidly expanding technical field in which it operates. Seven principal organizations, of which four are linked together by the Radio Industry Council, deal with various aspects of our industry, and in all cases there is a strong technical committee structure to discuss on an inter-company basis such matters as standardization, safety precautions and codes of good practice. Many of our members are prominent in these committees, and the way in which these work, both within themselves and with other bodies, such as the British Standards Institution and our own Institution, is a tribute to our professional standards.

Whilst there are some independent companies outside these organizations, I think it is true to say that on important issues it is possible to obtain the view of the whole of the industry.

Technical Manpower Requirements.—With the loss of our oversea investment in two world wars, and the scarcity of our natural resources, the future of this country depends above all on our ability to develop industries which will enable us to

Here we might usefully study the situation in Russia, where, in Moscow University, nearly 50% of the undergraduates in science and engineering are women, and the staff includes a number of women professors of engineering. Women engineers are to be found throughout Russian industry carrying technical and executive responsibility on an equal basis with men, while in the wider sphere it is claimed that 53% of college-trained specialists and 66% of people with a specialized secondary education are women. If it is correct that technical and administrative ability is as frequently present in women as in men—and there does not seem to be any evidence to the contrary—then the Russian experiment multiplies their potential yield by a factor of two, a much higher factor than is possible by other means.

To sum up: we must widen as far as possible the sources from which scientific and technological students can be drawn, and make sure that teachers, buildings and equipment are available for their education to the highest possible standards. Whatever we do, we should not compromise on our standards, as in the long run quality is more important than quantity.

New Materials and Techniques

It is generally true that research and development work in any field sooner or later meets limitations in the properties of the materials available, and this is very true of electronics. The post-war years have seen the development of a number of new devices derived from research into materials, and below are a few examples, all of which are associated with solid-state work.

Semi-Conductors.—The basic physics was outlined with great clarity by Mr. C. W. Oatley in his Chairman's Address to this Section two years ago, when he introduced us to the holes and electrons which carry current through a semi-conductor.

From a manufacturing viewpoint, artificially produced semi-conductors present a problem in *controlled impurity* as distinct from the classical chemical problem of *controlled purity.* In production, the basic material must first be purified to a degree far beyond normal manufacturing tolerances, and to this pure material must be added microscopic amounts, of the order of one part in 10^6, of an equally 'pure' impurity. Many ingenious methods are being developed for growing pure crystals and subsequently treating them by 'zone refining', and electronic methods of measuring their purity are available, but it is a process which requires pharmaceutical standards of cleanliness and a fundamentally different outlook in the factory.

Most of our electronic work to date has been concerned with electron flow in a vacuum—the so-called free electron—and in devices, such as transistors, making use of semi-conductors we are replacing this vacuum by an artificially produced crystal lattice. The material is difficult to control and inspect for flaws and variations, and as these are inherent in the medium it provides a problem likely to be with us for some considerable time.

The manufacturing problem lies in mass production to consistent tolerances and at a process cost which will take advantage of the low material cost. Discussion on whether transistors will entirely replace valves is I feel rather pointless—both devices have their place and are likely always to exist side by side, with the transistor taking over where small size, low power consumption and long life are of importance.

Phosphors.—Research work is being directed at producing more efficient phosphors with wider ranges of stimulation and emission radiation and of persistence. One objective is a photocell sensitive to nuclear radiation which will provide a radiation detector of simple design. Among other objectives are display devices for colour television and phosphors which, under ultraviolet stimulation or under the influence of a high-frequency field, can give us more efficient sources of artificial light. The stimulation of light from a phosphor by an alternating electric field, or electroluminescence (Fig. 2), is already being used for low-level decorative lighting, and work is going on in America using the combination of a light-sensitive phosphor and an electroluminescent panel to form a light amplifier for increasing the brightness of an image. This development should have many applications such as in television and in medical diagnostic work for improving the efficiency of fluoroscopy.

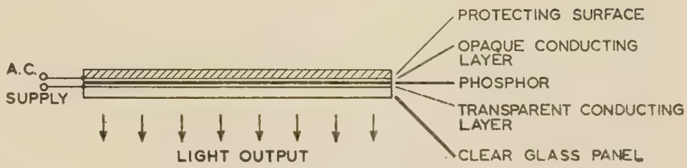

Fig. 2.—Section of electroluminescent panel.

Ferrites.—Chemically, ferrites have the formula MFe_2O_4, where M is a divalent metal. Physically they have the characteristics of ceramics and in fact are spoken of in manufacture as 'black ceramics'. Their great advantage over earlier materials is their low high-frequency losses. Three of their applications derive from the forms of hysteresis loop which they can be made to exhibit:

(*a*) Material with high permeability and a 'thin' hysteresis loop has extensive application as a core for high-frequency coils.

(*b*) Material with a high remanence and a 'fat' hysteresis loop is suitable for use as a permanent magnet which yet has low losses in a radio-frequency field.

(*c*) Material with a hysteresis loop with steep sides and roughly rectangular in shape leads to what are known as 'square loop' applications for use as a 'memory' in computer work.

Other applications of ferrites arise from electron spin and are of particular value in microwave circuits (Fig. 3).

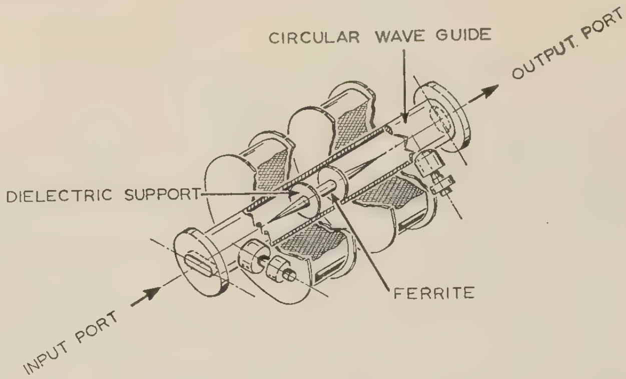

Fig. 3.—Ferrite microwave switching device using Faraday rotation effect.

At a Convention on Ferrites to be held here in a few weeks' time, some 50 papers and contributions from a number of countries will be presented.

I think it is no exaggeration to say that transistors and ferrites can each in their own field become as important to the electronic industry of the future as is the valve at the present time.

Miniaturization and Potted and Printed Circuits.—This is often described as a new manufacturing technique, although in fact it is a logical production step resulting from the latest component developments and the desire to transfer manual assembly and wiring operations into the field of fabrication. In all manufacturing processes there is a constant search for reduction of size so that advantage may be taken of the resulting lower

material costs and the lower factory costs resulting from the smaller weight and size. One example of this steady evolution is given by the successive reduction in sizes of receiving valves from the pre-war types to miniature and sub-miniature designs. Another lies in coils for carrier-current telephony, where the use of the new ferrite materials in place of the earlier dust cores makes a spectacular reduction in size.

In a production line, automatic machines may be used for inserting components, fed from a 'bandolier' roll, into a printed-circuit wafer automatically moved from one machine position to another on the transfer principle. Although only experimental lines are running on this basis, it could result in considerable reduction in manufacturing cost, provided that the production run was long and the design sufficiently stable to justify the high capital investment in the assembly line.

There are two main objectives in these manufacturing principles. The first is to reduce cost and improve uniformity and reliability on large-quantity production runs, and the second —which particularly applies to potting—is to produce a small and robust assembly where these factors are important as in certain military applications. These techniques involve a new principle of service and maintenance in which whole units are scrapped and replaced instead of faults being located and components replaced. Several manufacturers are experimenting with printed circuits and automatic assembly for domestic receivers, but it is not yet clear to what extent this development will be applicable to the manufacturing quantities with which we are concerned, since it is obvious that the savings increase with the length of the production run.

A new problem which arises from miniaturization is the removal of heat from the assembly, since there is a smaller overall cooling surface, and it is here that the transistor shows to great advantage over the valve through its much lower power consumption. It is my view that the real advantages of miniaturization and printed circuits will show up most clearly in those cases where a combination of transistors and ferrite materials can be used.

The Impact of New Materials and Techniques on a Number of Electronic Applications

Radio.—Transistors and ferrites have obvious application to mobile and portable apparatus where the saving of battery weight or the use of an already available low-voltage supply is attractive—portable receivers, tape recorders and record players, and car radio are typical examples. The development of frequency modulation in this country and elsewhere in the world throws great emphasis on the need for a transistor device suitable for the higher frequencies, and research is being directed to a solution of this problem.

Television.—Television is heading for a place in every home and for an important part of our export business. The main trend is an increase in the size of the picture—a development associated entirely with the receiving apparatus and its economics, unlike colour television which introduces problems of transmission and the cost of colour programmes. The sizes of the cathode-ray tubes for television receivers have been steadily growing from the 9 in and 12 in tubes with which we started to the present emphasis on 17 in and 21 in, but it is in my view doubtful whether it is practical to mass-produce tubes of larger sizes. Larger pictures are probably better produced by optical projection. Design developments have been directed at a reduction in the length of the tube, and hence a shallower cabinet, by scanning over a wider angle. Initially this was 70°, it is now moving to 90°, and the future may possibly see 120°. The high scanning power necessary would not be practicable without the use of ferrites, and this is a clear example of their application. Some research work has been aimed at reducing the length of the tube still further to form the flat tube often known as the 'picture on the wall' type. If this does come—and it can only be made possible by different scanning methods—ferrites and transistors will, I feel, have to play an important part.

The main unsolved problem in colour television is the display tube, and the only design at present in production is the 'shadow mask' type, in which the screen area is broken up into groups of three dots each with a different primary-colour phosphor, these dots being selectively stimulated, each colour by its own part of the electron beam. This construction is not only difficult and expensive to manufacture, but is inefficient in light production, and leads to thoughts of some new development which would enable the masking to be dispensed with. While this seems a remote possibility at the present time, it is a field in which further developments in solid-state phenomena might well provide a more elegant solution.

Industrial Television.—Aside from the purely entertainment value of television there are many other important applications of a professional and industrial nature which are finding increasing use. The early television pioneers had more interest in these industrial applications than in entertainment, so that the present trend is a return to their early intentions. For industrial use the apparatus can be considerably simplified and made cheaper and smaller. By suitable casing of the camera to accommodate the conditions of use, which may be extreme heat, high radiation levels, immersion in water or other liquids or high dust concentration, and the provision of suitable lens systems, which may be twin-lens, zoom-lens or even stereoscopic, we have a means of extending our vision to places which are either inaccessible or dangerous. This information may be presented in a number of different places at the same time, or alternatively several different views may be presented on adjacent screens. Testing and inspection applications using this technique in such important fields as atomic energy and jet aircraft will assist in increasing the pace of engineering development.

Underwater television is of unique application to salvage and rescue work, and in fishery research, while the addition of sight to sound in business communication may even lead to some decentralization from our overcrowded cities. In the professional field, television techniques hold out the hope of improving X-ray medical diagnosis, particularly if phosphor development can lead to an X-ray-sensitive television pick-up tube. Television is also of great assistance in microscopy, and, at the other end of the scale of dimensions, in astronomy. Recent work at one of our universities leads to the intriguing possibility of extending by television the range of the largest radio telescopes to 5 000 million light-years, which is about the distance light would travel during the time that the universe is thought to have existed.

The most important television application of all is to education and training, since it enables a teacher who is a master of his subject to present his knowledge simultaneously to an audience of millions, with close-up views of diagrams, experiments, films and other visual aids, in a way not possible by other means. Perhaps we have here the long-term solution to some of our educational problems.

Communication.—New developments in communication have most attractive applications in the switching, memorizing and routing problems in our telephone exchange network. The replacement of the human switchboard operator by the elaborate electromagnetic devices in our automatic exchanges has made possible a degree of subscriber density which would never have been possible by other means. Electronics, however, offers a great opportunity of improving the engineering efficiency of automatic exchanges, and the exchange of the future is likely

be mainly based on transistor and ferrite techniques associated th printed and potted circuits. Experimental work is already advanced on telephone exchanges using valves, and on the sis of the experience gained, this should not only avoid a eat deal of mechanical maintenance, but permit dialling direct er the whole of the trunking system. It is conceivable that in future we shall be able to dial from our own home telephone er a large part of the world via intercontinental links, using omarine cables with submerged repeaters. Forward-scatter d cable techniques will permit of world-wide television links. nger-term developments may make use of simpler cables with ilt-in transistor amplifiers or microwave guides.

Automation.—Transistors, ferrites and phosphors will play an portant role in automation—a new word for an old manu-turing principle. While the real meaning of the term is chanization of the 'closed loop' and consequently self-correct- type, in the mind of the public it is associated with any chanization which leads to improvement in production effi-ncy and hence regrouping of the labour force. Chemical plants d oil refineries already make extensive use of automation, but most potent application is likely to be in the metal fabrication lustries, in which we can include radio, where it is estimated it 15–20% of factories would benefit. While overall it is likely to have an immediate effect on more than about 10% employed persons, it involves the labour unions, and will l for diplomatic and tactful handling in its introduction. That s of benefit to mankind is in no doubt. In the last hundred ars, hours of work are estimated to have been reduced by % and the value of wages to have increased by 250% owing the development of production machinery, and further provements in standards of living lie ahead. Allied to reased availability of power from atomic energy, these velopments may bring back to the more highly industralized tions some of the advantages which they lost through the ablishment of the older machine industries, like textile manu-ture, in parts of the world where labour is less expensive, and must make sure in this country that we maintain a leading sition. In 1952 the kilowatt-hours per man-hour available in United Kingdom were 2·4 as compared with 8·6 in the nited States, and even to hold this ratio, let alone increase it, l need all our efforts. Electronics play an essential part in tomation owing to the vast range of physical quantities which capable of measurement and the flexibility of electrical ntrol. Another field of application lies in clerical work, where mputers should be able to take over a considerable proportion the work of the estimated two million clerks we have in this untry.

Atomic Energy.—Solid-state effects play an important role in many sensing devices that have been developed, and the sentation of this information, either in visible form or through production of controlling voltages, brings into use all the est developments of electronics. I think it is true to say that, thout the recent developments of electronic engineering, atomic wer, on which so much depends, would never have been possible. Even more important is its role in future power production with its attractive possibilities of deriving electrical energy direct from atomic fission and fusion effects without the intervention of the heat engine. Radioactive isotopes and their associated measuring instruments bring a new and powerful research tool to the agriculturalist, medical man and engineer, while high-energy particle accelerators and electronic devices are of increasing application to medical therapy.

Defence.—Miniaturization and the transistor and ferrites are of vital application to defence, since as speeds increase we have to depend more and more on automatic devices, and at the same time these higher speeds throw great emphasis on small size and weight. The guided-missile intercepter is a clear case in point. Infra-red devices give us sight in the dark and sense minute amounts of heat radiation, ultrasonic waves permit location of underwater objects, and the reflection properties of high-frequency radio waves enable us to locate invaders in the air. Valve techniques lead to large and complex apparatus, since a modern bomber carries equipment using 3000 valves and a cruiser uses 10000, the power supplies and heat dissipation alone being major problems. In future defence work, miniature components and automatic assembly will be a vital munition industry.

The Future

In this very broad review I have attempted the hazardous task of trying to look into the future, a task perhaps more dangerous in this technical field than any other. We are without doubt on the edge of a vast new area of exploration in sub-atomic engineering in which it is essential that we hold our own against the other industrial nations. Our research laboratories in the universities, Government establishments and industry have pioneered many of the most important fundamental discoveries, but so often we have failed to grasp the opportunity of putting them to practical use and profit. This new atomic field is eminently suited for the type of research in which we traditionally lead, and I feel that a special responsibility rests on the engineers of this Institution to ensure that we put these new discoveries to work with imagination and intelligence for the betterment of mankind. Never before in our history has it been so important nor has such supreme opportunity been offered. Technique is advancing so rapidly with the developments of sub-atomic physics that our production of power and its control and our means of communication and measurement offer a world of industrial exploration comparable to the geographical opportunities of past centuries. To-day history has been made by our Queen in opening Calder Hall, the first large-scale atomic power station in the world, and the bold plans for the series of atomic power stations of which this is only the beginning offer for future generations power on a scale beyond our ken. We must make sure that our lead is not wasted.

The risks are great, but so is the prize, and I have every confidence that with wise planning and sound engineering we in this country can make of our opportunities as much as did our forefathers a century ago.

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Jan. 1957

SHIP STABILIZATION: AUTOMATIC CONTROLS, COMPUTED AND IN PRACTICE

By J. BELL, M.Sc., Member.

(*The paper was received* 20*th September*, 1956. *It is based on the Measurement and Control Section Annual Lecture, which the author delivered on the* 8*th May*, 1956.)

SUMMARY

The paper relates in brief the development of the control apparatus by the author for the Denny–Brown type of ship stabilizer from the initial ON-OFF control, to the present proportional control using roll and roll-velocity functions. The use of roll acceleration and a feedback control are also discussed.

Predictions of the performance of a stabilizer by step-by-step and analogue methods are given, the functioning of the analogue computer is described and examples of results are included.

Some practical results from sea experience and a brief account of the stabilizing demonstration given during the lecture also appear.

The paper is complementary to one being presented before The Institution of Naval Architects, which deals more with the ship aspect of the subject.

(1) HISTORICAL

In the year 1875, W. Froude presented a paper to The Institution of Naval Architects describing a method of graphical integration by which, starting from a presumed train of waves and a given vessel's stability characteristic, the resultant motion of the vessel was computed. Such a method, while involving a great deal of labour in the calculations of a ship's acceleration, velocity and roll step by step, has the undoubted advantage that characteristics can be dealt with which are not easily soluble mathematically.

Following that paper were others dealing with stabilization against rolling by various means such as water tanks, moving weights, gyroscopes (large), water jets, and various ship forms. These have all been tried out over the intervening period but have not proved a success, to judge by the small number of practical installations. The remaining method uses keels or hydroplanes in one form or another.

Bilge Keels.—W. Froude early established the usefulness of bilge keels for roll damping, and they are normally fitted to vessels. The size of the keels is limited because they cause additional drag acting against the forward motion of the vessel and hence reduce the speed and increase the operating costs. They are most effective in reducing the maximum rolling, when synchronous conditions occur, i.e. when the natural period of the vessel and the waves coincide; at other times also they have a beneficial effect. Under the synchronous condition in a sea with one degree wave slope, a vessel not fitted with bilge keels might well roll $\pm 10°$ or 12°, while with bilge keels as normally fitted the roll would be about 6°.

Activated Fins.—Activated fins can be fitted to a vessel; these may consist of one or more fins of comparatively large dimensions and high aspect ratio, which are withdrawn into the vessel when not in use, or alternatively they can partially take the place of and be comparable in projecting length with the normal bilge keel. Such fins, for example, as proposed by the present author and fitted to one of H.M. cruisers, are of low aspect ratio and are not arranged to be withdrawn. A sufficient number of fins (8 in the case quoted) are fitted to make up the necessary effective control area.

The roll damping obtained with activated fins is of a different order from that obtainable with bilge keels or passive fins. Whereas with the latter the synchronous rolling amplitude of a vessel may be about 6 times the wave slope, with activated fins the Q-factor can be $0 \cdot 5$ or less; thus the damping obtainable is of the order of 12 times better than with the passive arrangement. This idea is not new—an early embodiment is described in Patent No. 19886 granted to Hiram S. Maxim in 1890. It is with installations of this kind that the present author's control apparatus has been designed and operated with some degree of success, and the computations of predicted performance for improved controls have been made.

During 1938, when on the scientific staff of the Admiralty, the author was first actively engaged on the problem of ship stabilization. Only two British vessels had at that time been fitted with a stabilizer of the Denny–Brown type, namely the *Isle of Sark* (cross-Channel steamer) and H.M.S. *Bittern* (a sloop). The stabilizer in its hydrodynamic aspects was described in a paper by Mr. (now Dr.) J. F. Allan,[11] and the present author contributed to the discussion.

(2) ON-OFF CONTROL

The first ship with which the author had to deal was the naval sloop H.M.S. *Bittern.* She was fitted with a system of ON-OFF control consisting of a small gyroscope mounted to be sensitive to roll velocity, and with or without so-called anticipation gear in operation, it made a contact for fin operation to counter the roll of the vessel either to port or starboard. The contact operated relays and solenoids controlling the hydraulic machinery to move the fins.

The number of strokes or fin movements in the action of the ON-OFF control, whether operated with 'anticipation' or without, tended to increase with light rolling. If heavy rolling occurs and maximum damping is required the fins must change attitude at each half-cycle of roll. Under conditions of light rolling, however, the operation is different and can be visualized as shown in Fig. 1. For the present purposes it is assumed that a vessel is resting on a wave slope such that a constant rolling torque is being applied to it and that this torque is, say, one-third of the available stabilizing torque, producing a rolling acceleration of $\ddot{\theta}$. The vessel is assumed to be initially at rest, but after a time t_1 it will have acquired a rolling velocity $\dot{\theta} = t_1\ddot{\theta}$. During this interval the fin motion has been initiated, and at time t_1 becomes fully operative (see Fig. 1).

The vessel now being subject to a torque of $\ddot{\theta}$ due to the wave slope and $-3\ddot{\theta}$ due to the fins, the resulting torque $-2\ddot{\theta}$ will reduce the rolling velocity $\dot{\theta}$ to zero and to a negative value in a further time t_2. The negative value results from the lags in centralizing the fins after the control signal has been cancelled. The sea only is now acting on the vessel and the motion and fin action are repeated all over again.

Mr. Bell is with Muirhead and Co., Ltd.

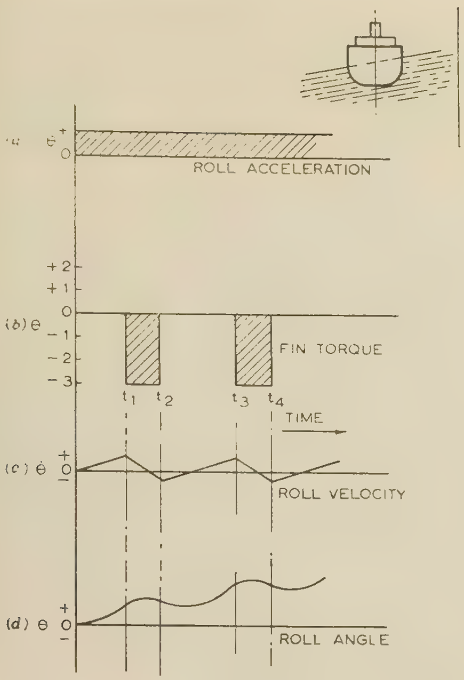


Fig. 1.—ON-OFF control: stabilizer action.
Small seas plus lag in response.

:t will be appreciated that these curves are qualitative only: s clear, for example, that the roll angle of the vessel (*d*) will odify the acceleration (*a*) since the relative angle between p's deck plane and sea surface has altered during the time der consideration.

:n the course of experiments on the *Bittern* after the time lags d been reduced to a practical minimum, various degrees of sitivity of the contact system on the velocity-responsive gyro re tried, and when the setting was too sensitive the familiar nting' phenomenon occurred, the movement of the vessel ng similar in character to curve (*d*) in Fig. 1 but continuing er the initial disturbing acceleration had ceased.

At the time (1939–1940) when this stage of the work was being ried out, 'hunting' was a well-known phenomenon in control paratus, as also was the remedy, namely a wider setting in ms of velocity or spacing of the gap between the operating ntacts of an ON-OFF control. In more modern terms of servo-chanism theory with a block diagram as in Fig. 2 (full lines

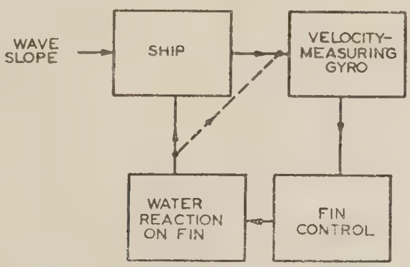


Fig. 2.—ON-OFF control: block diagram.

y), it can readily be seen that with an appropriate time lag nd the closed loop a continuous oscillation will occur whose quency is related to the frequency of response of the operating chanism rather than to the natural rolling period of the vessel. n practice it was found that the critical setting of the ON-OFF tacts was equivalent to approximately 0·5° per second rolling velocity. A stable setting was thus possible by which, if a greater roll than 1° amplitude with a 10 sec period occurred, a contact would be established and a fin movement result in a direction to diminish the roll. Smaller rolls than ±1° are ignored by this control.

Examples of stabilizing records obtained with the ON-OFF control are given in Figs. 3 and 4. Both were taken by the

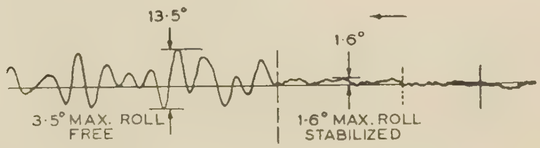


Fig. 3.—Gentle swell: synchronous rolling.

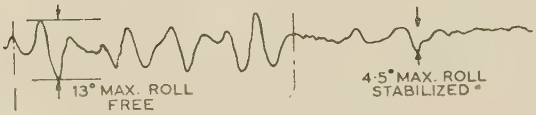


Fig. 4.—Confused sea: free rolling and stabilizing.

author on the *Bittern* in 1939 and clearly show in the stabilizing portion a roll in the period of the control gear of 2·5–3 sec and amplitude ±0·25°, representing a maximum roll velocity of the order of 0·6 deg/sec.

Fig. 3 was taken under conditions which are ideal for stabilizing, namely a gentle swell on the beam of the appropriate frequency giving approximately synchronous rolling of the vessel with the stabilizer not in action. The reduction of motion from 13·5° to 1·6° may be regarded as a highly satisfactory performance for the equipment in use, but the residual motion of short period due to fin 'hunting', while not distressing, is quite noticeable and does not meet the naval requirement of an ideal gunnery platform.

Fig. 4 shows the type of stabilizing obtained in a more rough and confused sea such as is frequently encountered in the English Channel. The reduction of rolling is still significant, but it is obvious that the power of the stabilizer is inadequate to hold the vessel against the larger sea forces encountered from time to time. Between these occasions of large wave motion, the vessel has practically no sinusoidal motion except that due to the ON-OFF 'hunting'.

The performance of the stabilizer with ON-OFF control, and with the time lag of all the stages of the control reduced to a minimum, is very good when considered as a roll damper only; under conditions of large acceleration of the vessel by the sea, however, this control suffers a disadvantage because of the time delay in its operation, in that a definite velocity of roll (0·5 deg/sec) must be built up before the contact is made, and a further increment of rolling velocity occurs during the interval before the fins are fully operative. If the condition then exists that the fin torque neutralizes only the acceleration being imposed on the vessel, the velocity already acquired may build up to a considerable roll angle before the wave passes the vessel, and the rolling energy thus built up has to be neutralized by a subsequent operation of the fins.

In the years 1939 to 1941 some 30 to 40 naval craft of the sloop and small destroyer (Hunt) classes were fitted with fin stabilizers having the control described, and reports were obtained from some of the vessels on the behaviour of the equipment. Three definite points arose out of these reports:

(*a*) That the large acceleration and short-period motion due to the ON-OFF control was a decided disadvantage.

(*b*) That when operating with a following sea, under which condition the rate of encounter of successive waves by the vessel is slow (i.e. the waves have a longer period than the rolling period of the vessel), the vessel tends to heel over to the wave slope while rolling in its own period. The control employing only a roll velocity-responsive gyro does nothing to counter this heel.

(*c*) The rapid and frequent fin movements caused mechanical vibration of the vessel and also electrical overloading of the auxiliary equipment.

To satisfy these objections and the new requirement it was clear that drastic alterations must be made in the control system and the main fin engine control, and the author was given wider powers to make the necessary proposals and to produce the gyro and servo control gear required.

(3) PROPORTIONAL CONTROL

A scheme was evolved which utilized a pendulous or vertical-keeping gyroscope for measuring roll angle and a velocity-responsive gyroscope with constraining springs for measuring the roll velocity.

Under conditions of heavy rolling as with the previous ON-OFF control the stabilizer had to be controlled as an active damping agency, opposing any motion of the vessel. This involved for highest efficiency the familiar question of anticipation of the velocity function in order to leave time for the fin engine to change the attitude of the fins at the moment of zero velocity. The roll angle lags behind the roll velocity function by 90°, so that adding velocity and angle functions together would produce more lag; but if the sign of the roll angle were reversed, a lead would result from the addition.

One combined control function used, therefore, and called 'beam sea' control, consisted of the roll velocity and -10% of roll angle, the amount of 10% being required to give the necessary time lead to the signal.

To provide for the following-sea control, the roll velocity function was combined with roll angle in roughly equal amounts. Thus both damping and vertical keeping features are present, but under conditions of heavy rolling the damping efficiency of this system may be expected to be lower than that of the beam-sea control because of the time lag in the change of attitude of the fins after the end of the roll (moment of zero velocity).

The practical scheme of carrying into effect the new control was achieved by the use of components and servo units already developed by the author and his colleagues for fire control and similar purposes. Magslips were used to accept the small mechanical signals from the gyros and add these together algebraically, and the power unit, a hydraulic servo of about $\frac{1}{4}$ h.p. output, responded with a lag not exceeding 0·2 sec. The arrangement is shown diagrammatically in Fig. 5, in which it can be seen that either of the two Magslips linked to the vertical-keeping gyro can be used by means of the beam-sea/following-sea change-over switch. The hydraulic unit drives into a gear box with appropriate reduction gear and operates a cam, limited in movement to $\pm 150°$. The cam surface consists of two opposite arcs struck from the centre but with differing radii and two transfer surfaces almost flat, connecting the arcs.

With zero signal from both gyros and with the Magslips and hydraulic relay unit in correct functional adjustment, the cam follower rollers lie in the centre of the transfer sections. Thus a signal in either sense will cause the cam to rotate, driving the follower and causing it to move the pilot-valve control arm (operating the main fin engines) with a movement proportional to the combined Magslip signal until the rollers reach the cylindrical surfaces of the cam; further Magslip signal causes additional rotation of the cam but no further movement of the pilot-valve control arm.

Fig. 6 shows the gyro control unit containing the two low-speed

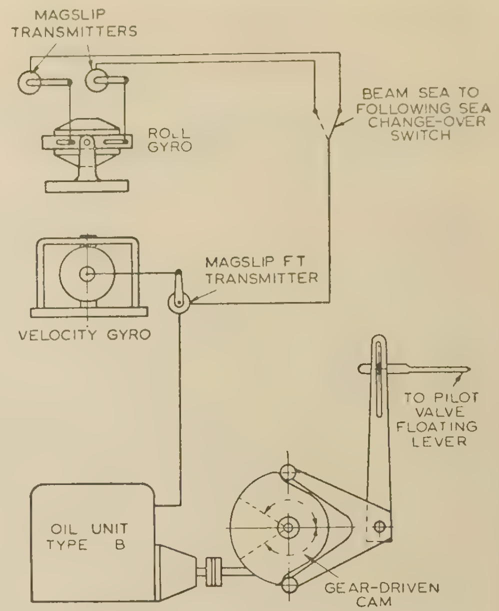


Fig. 5.—A.R.L. continuous control: schematic.

(3 000 r.p.m.) gyros designed for this duty, and also the Magslips mechanically linked to the gyros. A dashpot is coupled to the velocity-sensitive gyro to damp out any tendency to oscillate which might occur owing to ship vibrations.

Fig. 7 shows the combination of functions for the two controls, the shaded curves representing in each case the signal from the control gear to the fins. After allowing for the lag in the fin engine itself, the fins will operate at the corect point, as shown in the beam-sea diagram, and lag by an appropriate amount, as seen in the following-sea diagram.

(4) STABILITY CONSIDERATIONS

A ship is designed to be inherently stable, and its rolling period varies from about 3 sec for vessels of about 100 tons to about 25 sec for 80 000 tons.

A special case of instability when using ON-OFF control has been described in Section 2, but this instability was due to the particular limitations of ON-OFF control and a very limited angle of forced rolling resulted. Another case of instability in one of the early ON-OFF controls is worthy of note: two velocity-sensitive gyros of different design were fitted in the stabilizer compartment as alternatives for control of the stabilizer. The position of the platform on which one was mounted had not been well chosen, one end of the platform being attached to the ship structure and the other supported from the casing housing the fins.

At the dock trials, during which the fins were moved by manually deflecting the gyro to make the circuit for fin operation the behaviour was normal, but at sea, steaming at normal speed when the gyro made contact due to its precessional motion in response to a ship movement, the fin responded and immediately the gyro broke circuit and in some cases made a reverse contact and the fin centred (or reversed); the action then started all over again in rapid succession.

This occurred during the early part of the war, and time was no available for a thorough investigation, but it was definitel established that the effect observed was due to the tilting of th gyro platform when the forces set up by the reaction of the wate on the deflected fins caused a small flexure of the supportin structure. The block diagram for this oscillation is shown i

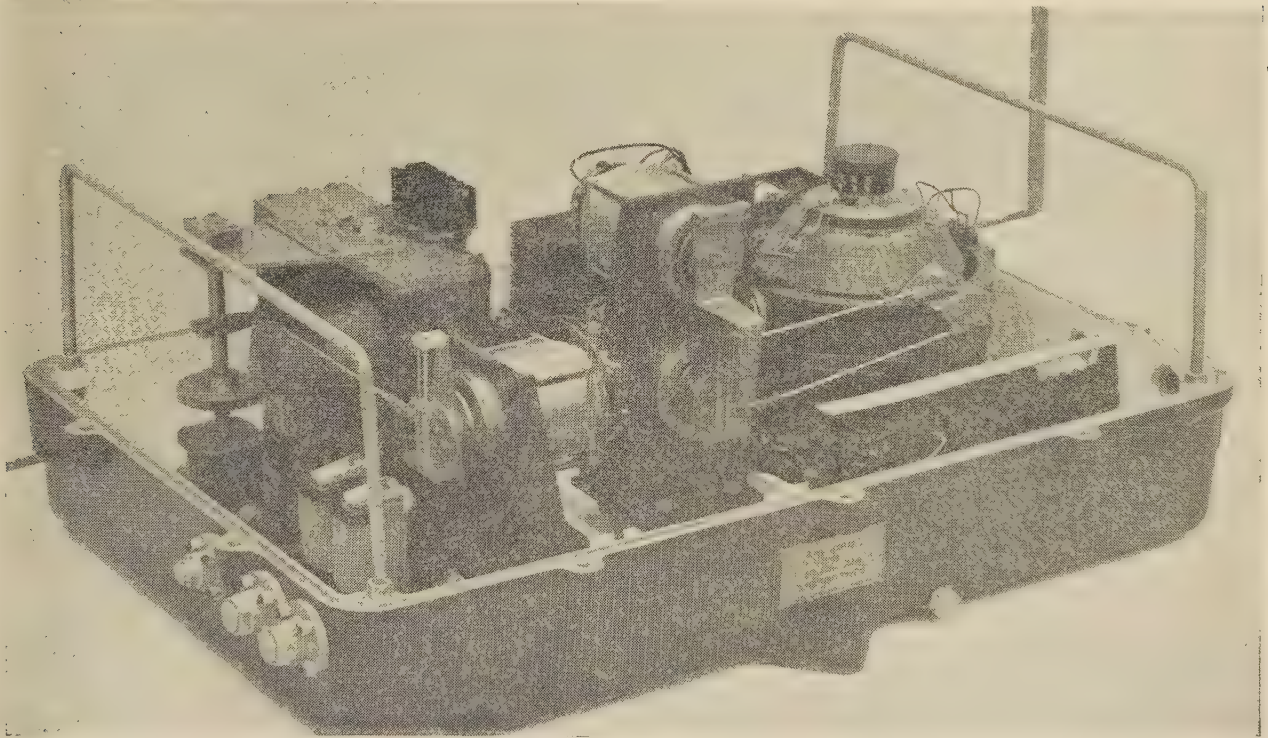


Fig. 6.—Gyro unit with cover removed.

2, if the closed loop is made following the dotted line and the uence of the inertia of the ship is removed from the chain of nts. The behaviour of the gyro, which was mounted on a cket securely attached to an athwartships bulkhead, was free n the rapid feedback oscillation and was quite normal.

nother case of incipient instability arises when acceleration ntroduced as a control function. When roll velocity or roll le is used as the control function a time delay due to the tia of the vessel is introduced between the application of the ilizing force and the appearance of this force in the control; a acceleration control, however, there is no delay. Thus a e slope may act on the vessel, the acceleration control onds, initiates the fin movement to cancel the wave force so reduce the net acceleration on the vessel to zero. This nediately cancels the acceleration signal and the fins go back ero. With a normal period of time lag in the operation in the a steady oscillation in their operating frequency can result. is obvious, however, that the use of acceleration as a control ction is desirable because it can be measured before the ocity or roll angle can be, and this time advance in fin movet gained should enable the stabilizer to keep the rolling vement of the vessel smaller, provided that the problems of ability can be solved.

(5) COMPENSATED CONTROL*

control scheme was devised (see Fig. 8) utilizing roll, roll ocity and acceleration, and also another function which may ailed fin 'wipe out' or feedback; this last is a control which is cted into the system and cancels the control due to the leration of the fins on the vessel and is derived from the ngle.

he mathematical expression involved in the movement of a el under the freely rolling condition is the same as for any

* Patent applied for.

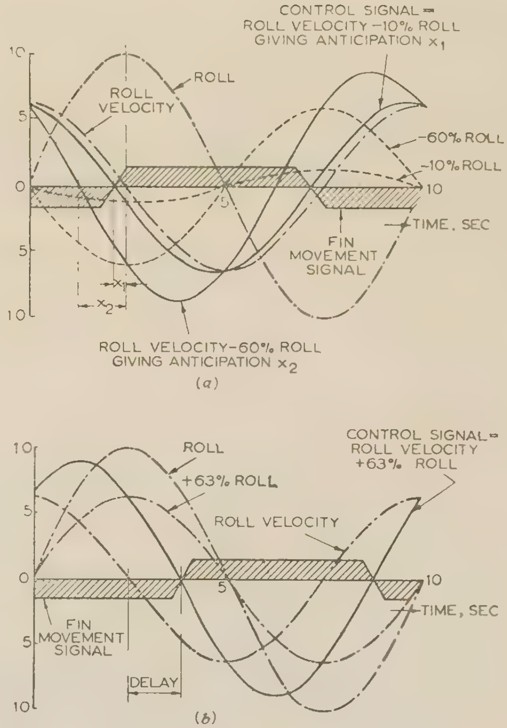


Fig. 7.—Control functions: A.R.L. continuous control.

(a) Beam-sea control.
(b) Following-sea control.

body having an approximately simple harmonic motion with damping, namely

$$a\frac{d^2\theta}{dt^2} + b\frac{d\theta}{dt} + c\theta = 0 \quad . \quad . \quad . \quad . \quad . \quad (1)$$

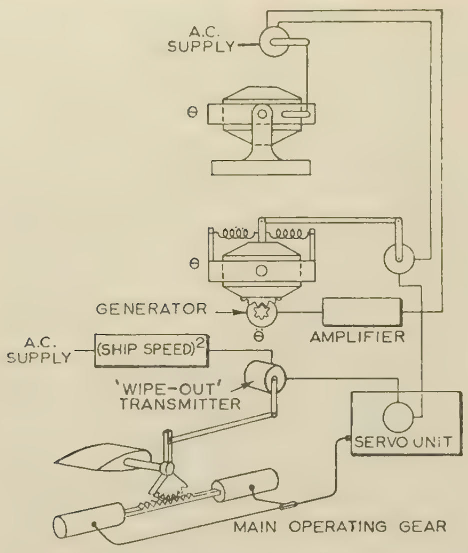


Fig. 8.—Compensated control: diagrammatic scheme.

This is an equation of torques or rolling forces, a, b and c being the constants of inertia, damping and righting moment respectively; if being disturbed by the sea the sea slope or acceleration, W should appear on the right-hand side instead of zero.

To introduce the stabilizer into the equation the control expression must be found and multiplied by the fin response. The measurements of roll and fin feedback are instantaneous, but the velocity and acceleration involve small time-constants due to the inertia of the velocity gyro and differentiating means. The fin-operating engine also has a time lag, and these three parts of the expression are therefore written with exponential functions. Thus the whole equation is as follows, the quantity S representing the stabilizing force at any moment, and α, β and γ being the time-constants referred to:

$$a\frac{d^2\theta}{dt^2} + b\frac{d\theta}{dt} + c\theta = \Big\{K_1[1 - \varepsilon^{-t/(\alpha+\beta)}]\frac{d^2\theta}{dt^2} + K_2(1 - \varepsilon^{-1/\alpha t})\frac{d\theta}{dt} + K_3\theta + K_4S\Big\}K_5(1 - \varepsilon^{-t/\gamma}) - W \quad . \quad (2)$$

(6) COMPUTING

Step-by-step method.—Previous work on the examination o the proportional control and its effect in countering various se disturbances had been carried out by the rather tedious an limited method, somewhat similar to Froude's original method of calculating step by step the motion of the vessel, and deducin at each step the control operative in the next. Fig. 9 shows th

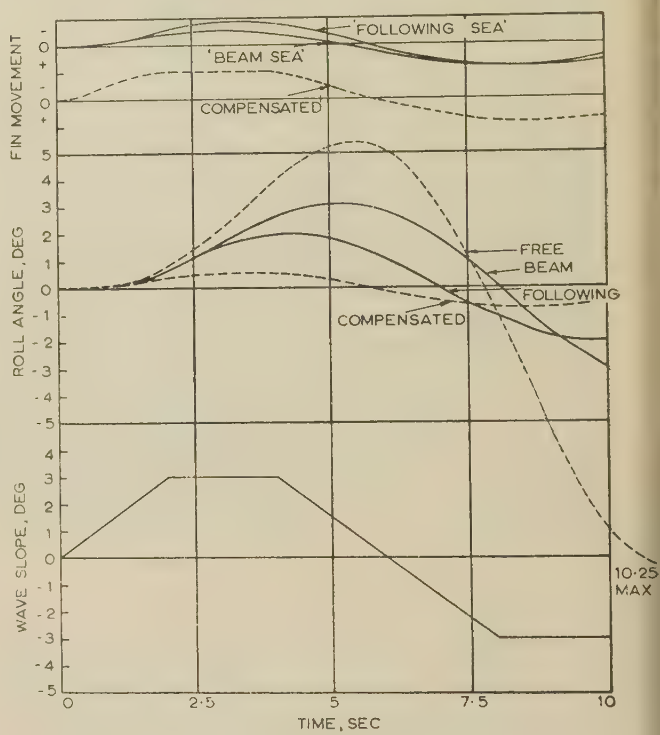


Fig. 9.—Ship stabilization compensated control proposals.

type of results obtained when plotted, the advantage in this cas of the new compensated control being clearly seen.

To cover the range of possible control parameters and als the variations of input or sea disturbances, it was obvious tha some more rapid method was essential. Analogue computin was therefore carried out using the facilities at the Nationa

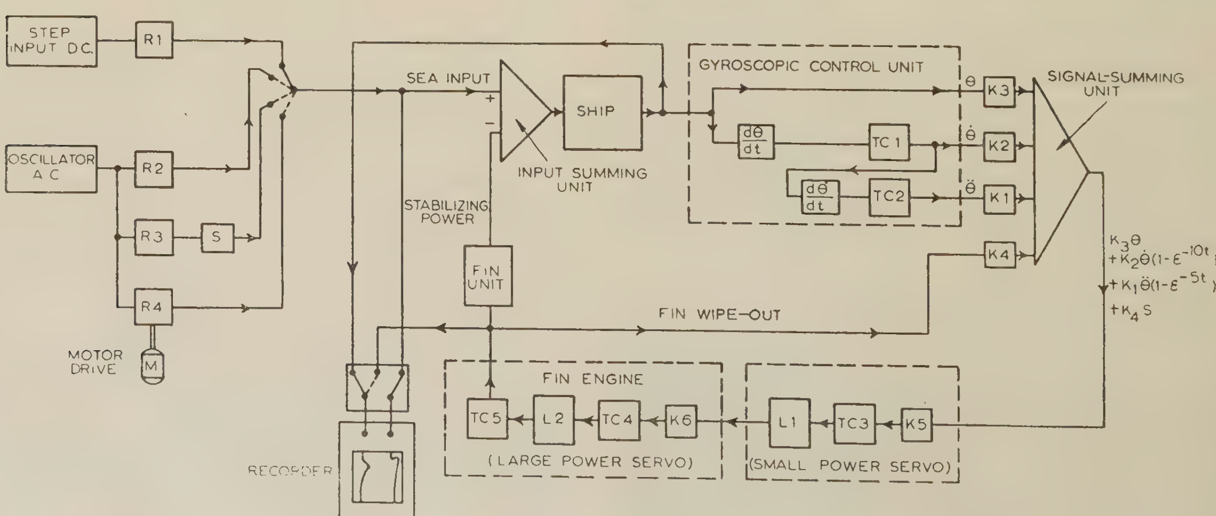


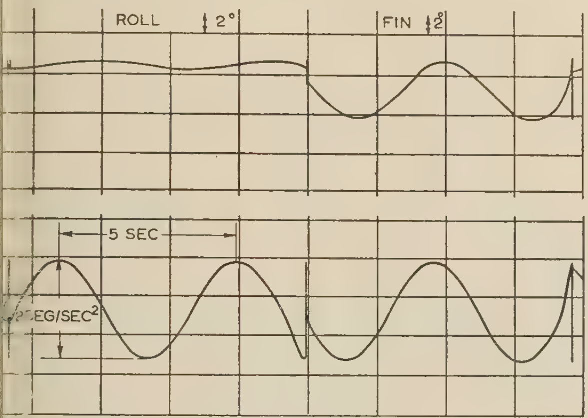
Fig. 10.—Block diagram of analogue computer.

ısical Laboratory. Some 300 runs were made under varying ıditions, and the results were recorded for future reference.

ınalogue method.—In this analogue computer all the quantities ɜ electrical, the ship being represented by an oscillating circuit ing appropriate damping. For some of the runs, as a matter onvenience, the frequency of oscillation was of the order of /s instead of the 0·1 c/s of the actual ship, and the other stants were adjusted to correspond.

ig. 10 is a block diagram showing the input side arranged step, sinusoidal, single-wave and increasing sinusoidal disances, and also the summing amplifiers, time-constants), constants (K) and limiters (L) representing the ship, ing unit and hydraulic servos.

ypical recordings from the computer are given in Figs. 11–14.



11.—Ship response and fin movement (equivalent list) for a continuous sinusoidal input.

(*a*) Compensated control.
$\theta = 4°$.
$\dot{\theta} = 2 \cdot 5$ deg/sec.
$\ddot{\theta} = 1 \cdot 4$ deg/sec.²
Feedback, 83%.

(*b*) Input: ± 1 deg/sec² peak, 5 sec period.

Fin power, 3° list. Fin operating time, 1 sec.

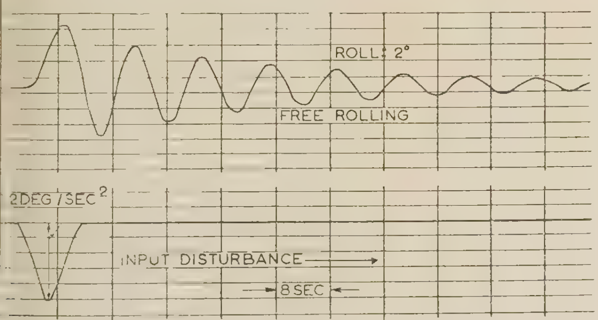


Fig. 12.—Ship response (free rolling) for a single sinusoidal disturbance, 2 deg/sec².

se show input accelerations of various types, 1°/sec² repreing a wave slope of 2·5°.

Fig. 11, a 5 sec sea of $\pm 2\frac{1}{2}°$ slope, which would have induced 1° roll in a 10 sec ship, is stabilized to $\pm 0 \cdot 4°$.

ig. 12 gives the effect of a single wave of 5° maximum slope 10 sec period, inducing free rolling of $\pm 7°$ and gradually ying with the natural decrement due to bilge keels and hull.

ig. 13 shows the stabilizer, whose maximum capability repre-

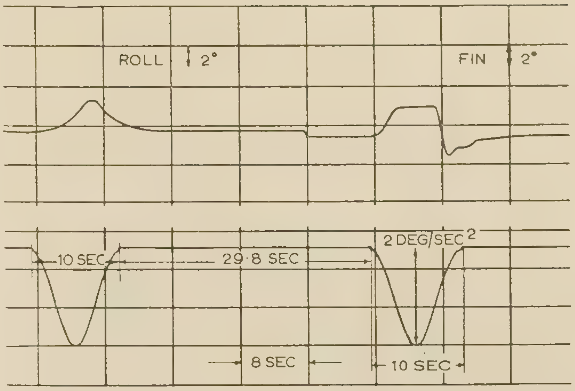


Fig. 13.—Ship response and fin movement (equivalent list) for a sinusoidal disturbance.

(*a*) Compensated control.
$\theta = 3°$.
$\dot{\theta} = 1$ deg/sec.
$\ddot{\theta} = 1$ deg/sec.²
Feedback, 94%.

(*b*) Input disturbances, 2 deg/sec².

Fin power, 3° list. Fin operating time, 2 sec.

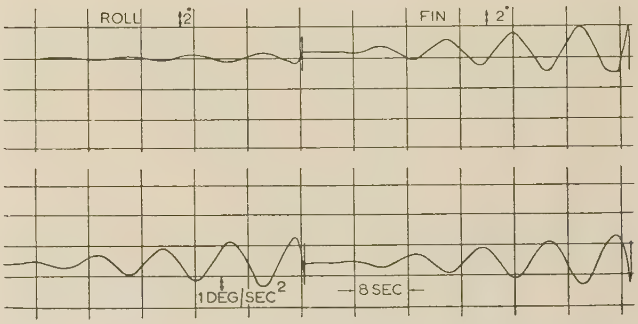


Fig. 14.—Ship response and fin movement (equivalent list) for an increasing sinusoidal disturbance.

(*a*) Compensated control.
$\theta = 3°$.
$\dot{\theta} = 1$ deg/sec.
$\ddot{\theta} = 1$ deg/sec.²
Feedback, 94%.

(*b*) Input, 0 to $\pm$ 1·6 deg/sec.²

Fin power, 3° list. Fin operating time, 2 sec.

sents only the equivalent of 3° wave slope, in action and quelling the same disturbance. The vessel swings only once about 3° from the vertical.

Fig. 14 illustrates waves of increasing slope up to 4°, and with the stabilizer in action the ship's motion is kept within $\pm 0 \cdot 6°$.

The repeated records in Figs. 11, 13 and 14 shows the fin action corresponding to the ship roll in the earlier part of the record. It will be observed that the fin reaches saturation in Fig. 13 and towards the end of the record in Fig. 14.

(7) DEMONSTRATIONS

Fig. 15 (a record taken on H.M.Y. *Britannia*) is included to show the type of results obtainable in moderately rough seas with the present proportional control.

On the occasion of the lecture this control was demonstrated on a dummy ship consisting of a rolling table about 4 feet square tuned to about 5 sec period and having a Q-factor of about 6.

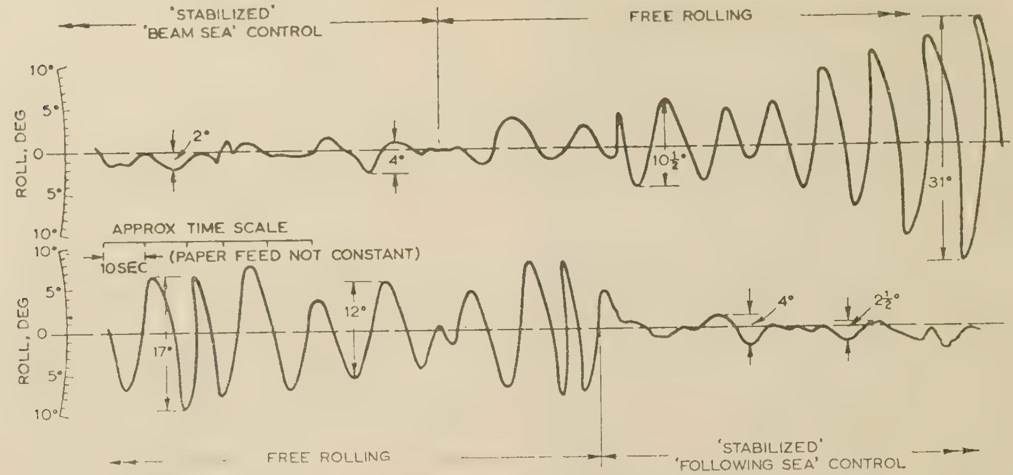


Fig. 15.—H.M.Y. *Britannia*: extracts from records taken on stabilizer trials, February, 1954, showing ship movement.

The table was caused to roll by the reaction of a rotating air jet and was then stabilized successfully by the operation of the control gear.

(8) ACKNOWLEDGMENTS

The author esteems it a privilege to have been associated with the development of the stabilizer, and in particular to have been instrumental, during the time of stress in the last war, in evolving a method of control which redeemed the system from disrepute and assured its present success in application to many naval vessels, including the Royal Yacht *Britannia* and upwards of 50 passenger vessels to the present time.

Thanks are due to Messrs. Muirhead and Co., Ltd., for providing the facilities for the recent developments and for their permission to the author to write this paper for publication; to Mr. R. H. Tizard for his co-operation in the analogue computing; and to the author's assistants, Mr. A. E. W. Hibbitt and Mr. J. H. Batchelor, of Muirhead and Co.

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368281	Minorsky N.	Improvements in stabilizing apparatus for ships and aircraft (control gear).	1929
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471220	Wallace, W. (Denny–Brown)	Improvements in and relating to anti-rolling apparatus for ships (anticipation control).	1938
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580982	Bell, J.	Improvements in and relating to means for stabilizing ships (multi fins non-retractable).	1944
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396.677

Paper No. 2216 R
Jan. 1957

THE RADIATING PROPERTIES OF END-FIRE AERIALS

By J. BROWN, M.A., Ph.D., Associate Member, and J. O. SPECTOR, Ph.D.

(*The paper was first received* 22*nd June, and in revised form* 10*th August*, 1956.)

SUMMARY

n end-fire aerial, such as a Yagi or a dielectric-rod aerial, can be ·ded as a structure which supports the dipole type of surface wave. ation occurs at the end of the aerial, the radiation pattern being rmined by the transverse field distribution of the surface wave. It own that this leads to a simple relation between the beam width e pattern and the ratio of the wavelength of the surface wave to in free space. Further, this approach shows that the beam width s to a non-zero value as the aerial length increases, in contrast with ous theories which predict that it decreases to zero with increasing h. In applying the theory to practical aerials, allowance must be for the radiation which also occurs from the end of the aerial at h the surface wave is excited. Theoretical and experimental rns for several end-fire aerials are compared, and it is shown that agreement can be obtained.

LIST OF SYMBOLS

λ_0 = Free-space wavelength.
λ_g = Wavelength of the dipole-type mode travelling along a dielectric rod.
L = Length of rod.
θ = Angle with rod axis.
$g(\theta)$ = Amplitude radiation pattern.
A, C = Arbitrary constants.
$\psi = \dfrac{2\pi L}{\lambda_0}\left(\dfrac{\lambda_0}{\lambda_g} - \cos\theta\right)$.
E_y, E_ρ, E_ϕ = Components of electric field.
G, k_0, k_1 = Constants defined by eqns. (10), (7), (8) and (9), respectively.
β, β_0 = Phase-change coefficients of dipole wave on rod and plane free-space wave, respectively.
$(x), K_n(x)$ = Bessel functions.
ϵ_r = Relative permittivity of dielectric forming the rod.
a = Radius of the rod.
$F(\rho, \phi)$ = Amplitude distribution.
$u = \beta_0 a \sin\theta$.

(1) INTRODUCTION

n end-fire aerial is one in which the direction of maximum tion is parallel, or nearly parallel, to the major dimension e aerial. The radiation pattern is usually symmetric about direction and thus has equal beam widths in its two principal es. Common examples of end-fire aerials are the Yagi[1] and [2] at metre and decimetre wavelengths, and the dielectric dielectric tube[3] and certain types of waveguide slot arrays[4] ntimetre wavelengths. End-fire aerials do not by themselves very narrow beams but can be used to build broadside arrays ve any desired beam width. In general, an end-fire aerial ss bulky than a broadside aerial with a similar radiation rn, and is therefore often preferred in applications where space is restricted or where the reduction of wind resistance is an important consideration.

Considerable work has been done on each of the individual aerials referred to above, but apparently little attempt has been made to discuss the general mechanism of radiation. Such a general approach is discussed here, and although, for simplicity, detailed comparisons are made with only two types, the Yagi and the dielectric rod, similar considerations apply to each of the others. A summary of the known theory for these two types is given in the next Section.

(2) SUMMARY OF EXISTING THEORIES

(2.1) The Yagi Aerial

As the oldest of the end-fire aerials, the Yagi has received most attention, and considerable theoretical information of a general nature is available. The normal arrangement comprises a driven half-wave or folded dipole, backed by a reflector and having a number of directors, as shown in Fig. 1. The driven element

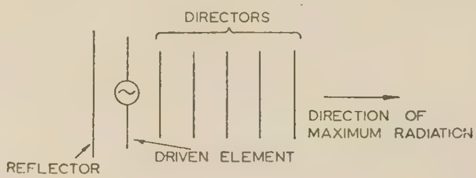

Fig. 1.—The Yagi aerial.

excites currents in the reflector and directors, whose amplitudes and phases can be calculated from the mutual impedances between the elements. The radiation pattern is then obtained by summing the contributions from each of these currents in the same way as for an array of driven elements. The analysis involved is described in detail by Pidduck,[5] but, while the method is formally possible for any number of elements, the computation required becomes prohibitive for large numbers. Numerical results for small numbers have been obtained by Walkinshaw,[6] and the choice of current amplitudes and phases to give minimum beam width for a given number of elements has been established.[7]

R. A. Smith[1] has suggested that the physical action of the directors is to reduce the phase velocity of the wave travelling along the axis of the Yagi. This is equivalent to saying that the wave radiated by the driven element travels through a region of refractive index greater than unity, and that, since the variation of refractive index in this region depends on the dimensions and spacing of the directors, the possibility arises of selecting them to give the same net effect as a convergent lens. This argument is not immediately amenable to mathematical analysis, but, as will be seen in Section 3, the physical mechanism envisaged is similar to that used here.

(2.2) Mallach's Theory for Dielectric-Rod Aerials

According to Mallach,[8] the action of a dielectric-rod aerial is to establish a wave travelling in the direction of the rod axis

ttea contributions on papers published without being read at meetings are r consideration with a view to publication.
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with a phase velocity less than that of a plane wave in free space. This wave sets up a continuous array of radiating elements, which can be regarded as the limiting case of the array of discrete elements shown in Fig. 2. The elements are assumed to radiate

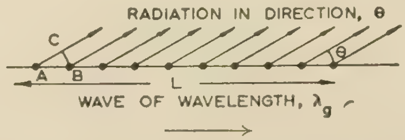

Fig. 2.—Array of elementary radiators equivalent to dielectric rod.
The phase difference between contributions from successive radiators is

$$\frac{2\pi AB}{\lambda_g} - \frac{2\pi AC}{\lambda_0} = 2\pi AB\left(\frac{1}{\lambda_g} - \frac{\cos\theta}{\lambda_0}\right)$$

equal amounts of power, with phase difference between successive elements depending on the free-space wavelength, λ_0, and the wavelength, λ_g, of the guided wave travelling along the rod axis. Superposition of the fields radiated by the elements gives, in the limiting case when the element spacing tends to zero, the following expression for the field strength radiated in the direction making an angle θ with the rod axis:

$$g(\theta) = \frac{C \sin\left(\frac{\pi L}{\lambda_g} - \frac{\pi L \cos\theta}{\lambda_0}\right)}{\pi L\left(\frac{1}{\lambda_g} - \frac{\cos\theta}{\lambda_0}\right)} \quad . \quad . \quad . \quad . \quad (1)$$

where C depends on the total radiated power and the distance at which the radiated field strength is calculated, and L is the length of the rod. The wave which is propagated along the rod is a surface wave of the dipole type, and its phase velocity depends on the diameter and relative permittivity of the rod. Numerical values have been calculated by Chandler[9] and verified experimentally by Elsasser.[10]

The field patterns for the dipole wave are indicated in Fig. 3, and expressions for the transverse components of the electric

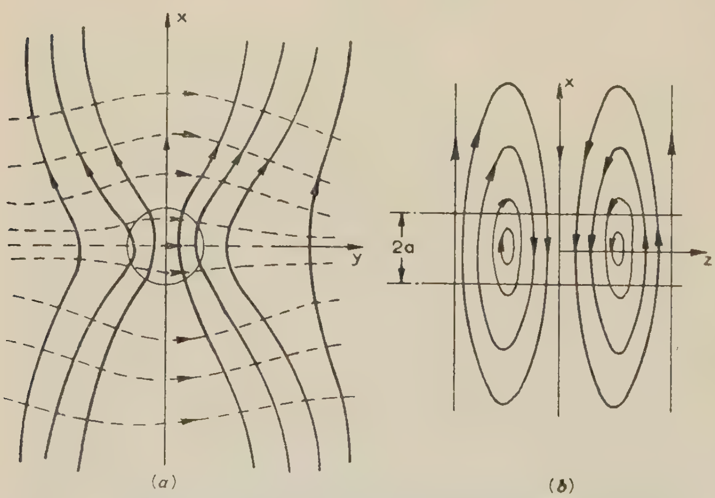

Fig. 3.—Sketch of the field distribution of a dipole wave travelling on a dielectric rod.
—— Electric field. ---- Magnetic field.
(a) Section normal to rod axis. (b) Section through rod axis.

field are given in Section 8.1. The amplitudes of the fields fall off quite rapidly with increasing distance from the rod axis, the rate of decay becoming larger if the rod diameter is increased. Since an increase of rod diameter also reduces the phase velocity of the wave, it follows that the effective extension of the wavefront from the rod axis depends inversely on the difference between the free-space velocity and that of the wave. A thin rod supports a spread-out wave travelling with nearly free-space phase velocity and a thick rod supports a wave concentrated near the surface of the rod with a phase velocity almost the same as that of a plane wave in the dielectric.

There are two objections to Mallach's theory; first, that it assumes a radiation mechanism which is physically impossible, and secondly, that the agreement with experimental results is poor, especially when the rod length exceeds a few wavelengths. The theory is therefore acceptable to neither the pure theoretician nor the practical engineer.

Objections on theoretical grounds are considered first. The postulated radiation mechanism supposes that energy is continuously radiated from the length of the rod, i.e. that the rod functions as a leaky waveguide. For the dipole-type surface wave, however, the only net flow of energy is in the direction of the rod axis, and a dielectric rod, far from being a leaky waveguide, can be used to guide energy over distances of many wavelengths with very little attenuation. The radiated wave cannot therefore arise as a result of leakage from the rod unless waves other than the dipole type exist. This by itself is not a complete condemnation of the theory, since a similar objection can be levelled at the well-established method for calculating the radiation pattern of a thin half-wave dipole. This rests on the assumption of a sinusoidal current distribution, which by itself cannot radiate power. For the half-wave dipole, a more complete analysis shows that a relatively small alteration to the current distribution suffices to produce a condition where power can be radiated, and that this change has no significant effect on the radiation pattern. No such complete analysis has been made for the dielectric rod, so there is no obvious justification for the assumption that fields other than the dipole mode do not exist.

The purely practical objection to Mallach's theory is that it fails to predict the observed dependence of the beam width of the aerial on the length of the rod. According to eqn. (1) the beam width should decrease steadily with increasing rod length, tending to zero as the length tends to infinity. In practice, the beam width may show an oscillatory dependence on length, reaching some 20° to 40° for lengths of a few wavelengths. Further increase in rod length has little or no effect on beam width. Yagi aerials show a similar behaviour.

(2.3) Horton and Watson's Theory for Dielectric Rod Aerials

Horton and Watson[12] apply the vector form of Green's theorem to relate the radiated field strength at any point to the distribution of electric and magnetic fields over the dielectric rod surface. If these fields are taken as those of the dipole wave, this approach gives the same results as Mallach's theory. In principle, the exact field distribution, including waves along the rod other than the dipole wave, can be used, and then the radiation pattern of the aerial can be calculated exactly. This theory can therefore be regarded as a rigorous form of Mallach's theory. The difficulty lies in the selection of the correct field distribution on the rod surface, and there appears to be no simple method of obtaining this.

(3) OUTLINE OF NEW THEORY

The basis of the theory proposed here is that the function of the guiding structure in an end-fire aerial is to distribute the energy supplied to the aerial over a radiating aperture. This can best be visualized by taking a specific end-fire aerial, e.g. a dielectric rod fed by a waveguide as shown in Fig. 4(*a*), and comparing it with a pyramidal electromagnetic horn [Fig. 4(*b*)], as is frequently done in discussions on rod aerials. The latter is a simple broadside aerial in which the radiating aperture is defined by the edge of the pyramid, i.e. ABCD. The flared portion serves to spread

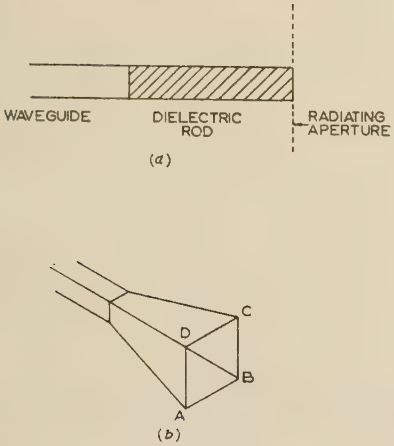

Fig. 4.—Comparison of waveguide-fed dielectric rod aerial with pyramidal horn.

(a) Waveguide-fed dielectric rod aerial.
(b) Pyramidal horn.

energy supplied by the waveguide feed over this aperture in a sonably uniform manner. The dielectric-rod aerial can be arded as a similar aerial in which the flare and the aperture CD are replaced, respectively, by the rod itself and by the ne at the end of the rod normal to the axis. According to this ture, the major function of the rod is to act as a convertor of H-mode, which is restricted to the confined space of the veguide, to the surface wave travelling along the rod, giving a ld distributed over an appreciable area of the radiating erture.

There are obvious dissimilarities between the two aerials shown Fig. 4, which may suggest that useful results are not likely be derived from the analogy. In fact, however, all the dif- ences, with one exception, are essentially differences in detail ly and are not such as to materially change the method of eration. The exception is the behaviour at the junction ween the waveguide feed and the guiding structure. In the rn, an incident wave in the guide excites a wave travelling twards in the flared portion, a reflected wave in the guide, and eactive field which is limited to a region close to the junction ween the guide and the flare. This reactive field does not olve any loss of power, and its only practical effect is that the ndwidth over which the reflected wave can be cancelled is ghtly restricted. The situation at the junction between the veguide and the dielectric rod is a little different: again, the cident wave generates a wave travelling along the guiding ucture—this time the rod—a reflected wave in the guide, and eactive field near the junction, but in addition there is a field rresponding to direct radiation from the position of the action. This last field represents a phenomenon not present the horn and arises because of the nature of the fields associated th a guiding structure such as a dielectric rod.[13] The field diated from the junction will be spread over a wide range of ections and can be assumed, in the absence of precise informa- n, to have a radiation pattern similar to one which would be med if the waveguide were allowed to radiate directly. The pportion of the incident radiation directly radiated from the c is obviously a factor which will exert a considerable influence the performance of the complete aerial. Calculations of the nching efficiency of structures used to excite surface waves ve been made,[14] but no figures are available for the arrangement olved in dielectric-rod aerials. One general feature of such tems is, however, quite clear, namely that the launching ciency will increase as the surface wave is more restricted to the vicinity of its guiding structure. The need to include the directly radiated wave in any estimate of the radiation pattern of the complete aerial is obvious. One reason for the poor agree- ment between previous theories and experimental results is the neglect of this wave.

Of the other dissimilarities between the horn and the dielectric rod, the most obvious is that the radiating aperture of the horn is clearly defined, whereas the dielectric rod will radiate through a region whose area depends on the detailed field distribution for the dipole wave travelling along the rod. The function of the horn flare is to spread the energy over an aperture of sufficient area to produce a radiated beam of the desired width. The wave travelling within the flare is spherical with a continuously expanding illuminated area: the dipole wave, on the other hand, maintains a constant illuminated area. A more exact equivalent to the horn appears, at first, to be a gradually tapered rod for which the illuminated area expands, but, as will be seen in Section 5, tapering the rod may introduce additional complica- tions. For horns of fixed aperture size, the radiation pattern does not depend to any marked extent on the flare length, unless this becomes so short that the phase varies appreciably over the aperture. For flare lengths exceeding the value required to keep the phase constant to within acceptable limits (usually about 20°), the radiation pattern becomes independent of this length. Increasing the length of a uniform dielectric rod diminishes the amplitude of the evanescent field which extends from the feed to the radiating aperture, and once the length is sufficient to make this evanescent field small compared with the dipole wave, further increases will not change the radiation from the aperture. So far the behaviour is similar to that of the horn. A com- plicating factor is the direct radiation from the waveguide feed, since changes in rod length vary the phase difference between the waves radiated from the feed and from the final aperture, thus causing changes in the observed radiation pattern.

The general conclusion is that the dielectric rod operates in a similar manner to an electromagnetic horn, with radiation occurring primarily at the free end of the rod. Some of the power fed to the aerial is radiated directly at the junction between the waveguide and the rod, and the complete radiation pattern must be obtained by summing the two contributions, making allowance for the phase difference between them. The first step towards a mathematical theory from which radiation patterns can be calculated is the derivation of the pattern corresponding to the field distribution for the dipole-type surface wave.

It is of interest to recall the considerable discussion in the past on the question whether an open-circuited 2-wire transmission line radiates from its end or continuously along its length. After many years, it was established that the former is correct.[15] The present problem is analogous in that a dielectric rod behaves as a transmission line: Mallach's theory rests on the incorrect assump- tion that radiation occurs continuously along the length of the rod, while the theory given here is based directly on the correct physical mechanism that radiation from a transmission line supporting its propagating mode can occur only at a discontinuity.

(4) COMPARISON OF THEORETICAL WITH EXPERIMENTAL RESULTS

(4.1) The Radiation Patterns of Long End-Fire Aerials

In Section 8.1, the field distribution for the dipole wave on a circular dielectric rod is examined and approximations are made on the assumption that the rod diameter is smaller than the operating wavelength. The dipole wave then has a wavelength just a little less than the free-space wavelength, and to a good degree of approximation the field distribution is linearly polarized. The magnitudes of both the electric and the magnetic field

strengths decay quite rapidly with increasing radial distance from the rod, so that the field in the aperture plane of Fig. 4(*a*) can be regarded as restricted to a finite area. The radiation from this aperture plane has been calculated assuming that the field distribution is that of the dipole wave. While this is not strictly true because of the excitation of evanescent waves which will travel back along the rod, the radiation pattern is not likely to be seriously in error. The neglect of the corresponding evanescent waves in the horn aperture makes no appreciable difference to the radiation pattern.

A selection of radiation patterns, calculated from eqn. (29) of Section 8.2 for polystyrene rods of various diameters, is shown in Fig. 5. The most striking features are the absence of side-

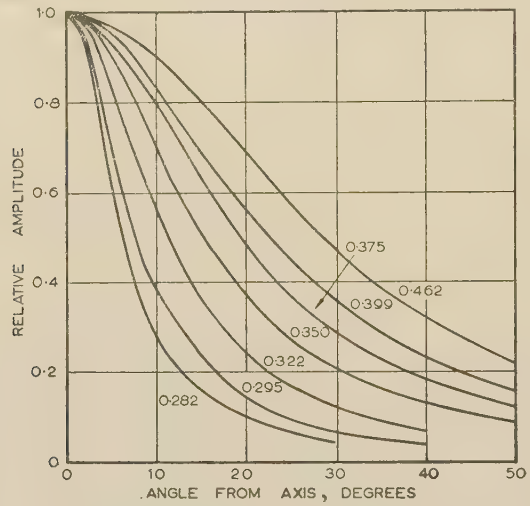

Fig. 5.—Theoretical radiation patterns for cylindrical polystyrene rods.
The parameter is the ratio of rod diameter to free-space wavelength.

lobes and the decrease in beam width which accompanies a decrease in the rod-diameter/wavelength ratio. The first of these is true only if the aperture field extends to an infinite distance from the rod, as assumed in the calculation, and is a property also possessed by the Gaussian distribution. The inclusion of the evanescent fields excited at the aperture will probably cause the appearance of small side-lobes, but investigation of this has not been attempted because of the more serious complication caused by direct radiation from the aerial feed. The bulk of the energy travels not in the rod but in the free space outside it, so that similar radiation patterns can be expected for any structure which supports a dipole-type wave. The beam width has accordingly been plotted in Fig. 6 in terms of the wavelengths in free space and along the guiding structure, and it is suggested that this curve is applicable to any end-fire aerial. The progressive decrease of beam width as the wavelength of the dipole wave tends to the free-space wavelength is to be expected on physical grounds. The smaller the difference between the wavelengths, the greater is the field extension of the dipole wave, thus increasing the size of the radiating aperture. The limiting case occurs when the guiding structure vanishes, leaving a plane wave which can be regarded as equivalent to a unidirectional beam.

The above results are not immediately applicable to a practical end-fire aerial because they neglect the direct radiation from the aerial feed. They do, however, suggest that there is a limit to the beam width as the aerial length is increased, and the patterns in Fig. 5 can be regarded as those for very long aerials for which the radiation from the feed can be neglected, since this will decrease in amplitude more rapidly than the wave travelling along the rod. This implies that the beam width of an end-fire aerial

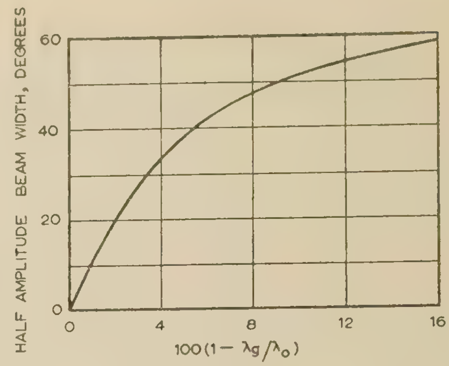

Fig. 6.—Theoretical dependence of half-amplitude beam width on the percentage difference between free-space and dipole-mode wavelengths.

should tend to a finite value depending on the effective phase velocity of the guided wave as the aerial length is increased.

(4.2) The Effect of Direct Radiation from the Aerial Feed

The radiation pattern of a practical end-fire aerial depends on the amount of energy radiated directly from the feed, as discussed in Section 3. Since no information is available on the fraction of the energy directly radiated, a trial-and-error approach has been adopted to determine the value which gives the best agreement with experimental results. Three radiation patterns due to Mallach[1] and Halliday and Kiely[16] for dielectric rods of different lengths, and with guided wavelengths approximately 10% different from the free-space wavelength, have been studied. In each case, it has been found that the best agreement between the theory given here and the experimental results is obtained when 6% of the energy is assumed to be directly radiated by the feed. The calculated and experimental patterns are shown in Fig. 7. The radiation from the free end of the rod is taken to be of the form shown in Fig. 5, and for the direct radiation, the dependence on direction is assumed to be so for an open-ended waveguide. The phase difference ψ between these two radiated beams for the direction making an angle θ with the rod axis is given by

$$\psi = \frac{2\pi L}{\lambda_0}\left(\frac{\lambda_0}{\lambda_g} - \cos\theta\right) \quad . \quad . \quad . \quad . \quad . \quad (2)$$

Eqn. (2) assumes that the radiated waves start off from the two ends of the rod with the phases of the dipole wave at these points. This is a plausible assumption, but it is possible that a detailed analysis would show the need for an additional constant term in the expression for ψ.

The agreement for the longest rod is surprisingly good in view of the approximations which have been made, and for the other two the general trend of the experimental patterns is followed quite closely by the theoretical curves. It is probable that the difference between the theoretical and experimental results arises from the evanescent fields in the aperture plane and from an incorrect radiation pattern being assumed for the direct radiation from the feed. The shape of the radiation patterns is such that a single parameter, such as half-amplitude beam width, conveys little information, but it is of interest to note that the smallest value occurs for the shortest rod. The reason for this is to be found in the variation of ψ with length and direction. For the shortest rod, ψ is 180° when θ has the value 28°, corresponding to the half-amplitude point if the feed radiation is neglected. In this case, the addition of the contribution from the feed has the effect of narrowing the beam. For the 4λ rod, ψ is 360° when θ is 28°, so that the feed contribution widens the beam, while

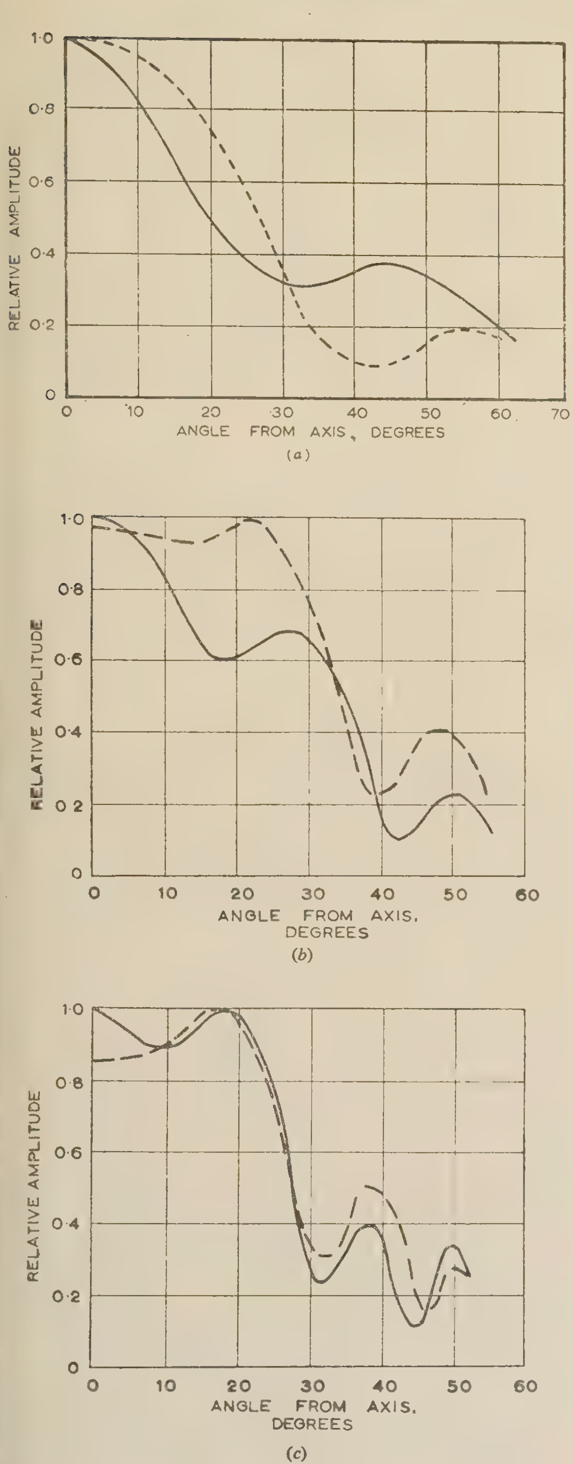

Fig. 7.—Radiation patterns for dielectric rods of diameter equal to $0{\cdot}46\lambda$.

— Experimental patterns from References 1 and 16.
– – Theoretical patterns calculated on the assumption that 6% of the total energy is directly radiated from the feed.
The lengths of the rods, in free-space wavelengths, are (a) 2, (b) 4, (c) 6.

the 6λ rod, ψ is 450° when θ is 28°, leaving the beam width tively unaltered.

the example discussed in Fig. 7, the surface wave is confined in a small distance from the rod axis and is launched with a high efficiency. For rods of smaller diameter, more of the incident power is radiated directly from the feed, and the radiation properties are therefore less dependent on the surface wave. The effect of this can be most clearly seen by considering the dependence of the beam width of a dielectric-rod aerial on the rod diameter. According to the theory given here, decreasing the rod diameter leads to an increase in the effective area of the radiating aperture at the end of the rod. This should cause a decrease in beam width, provided that all the available energy travels down the rod in the dipole mode. On the other hand, the increased spread of the field of this mode makes it more difficult to launch the energy in the mode, and an increase in the direct radiation from the feed results. This gives an increase in beam width for a decrease in rod diameter. Experimental results given by Halliday and Kiely show that, in practice, the beam width does increase with decreasing diameter for rods whose diameters are less than $0{\cdot}46\lambda$. The reduction in launching efficiency is thus the dominant factor.

(4.3) Yagi Aerials

There is a general similarity between a Yagi and a dielectric-rod aerial, the array of directors in the Yagi taking the place of the rod itself. The same radiation mechanism can be postulated, provided that the array of directors can support a pure travelling wave, without any radiation from its length. This has been examined experimentally[17] by constructing a long array formed by equispaced identical conducting rods, and short-circuiting the ends by large reflecting sheets to form a resonator. The phase velocity and attenuation of the wave travelling along the structure can be deduced from the resonant frequencies and Q-factors of the resonator. By this technique it has been established that a dipole wave can propagate along the structure without losing power by radiation, provided that the lengths of the conducting rods are less than half the free-space wavelength. It is, of course, already well known from experimental work on Yagi aerials that the directors should be less than half the free-space wavelength. From the resonator experiments, values for the phase velocity have been obtained for a wide range of conductor diameters, lengths and spacings, and from these values the beam width of a long Yagi can be predicted, using the curve of Fig. 6. A set of experimental results for long Yagis has been given by Fishenden and Wiblin,[11] up to 30 director elements being used. The beam width appears to tend to about 22° for large numbers of director elements: the corresponding theoretical figure, as predicted from Fig. 6, using the experimentally determined dipole-mode wavelength for a structure having the same dimensions as the Yagi array, is 25°. It might be expected that the theoretical figure should be smaller than the experimental, but there are two factors which could justify the opposite result. The first is the effect of direct radiation from the feed, which as seen in the previous Section can either increase or decrease the apparent beam width. The second is that the wavelength measurements on the Yagi-type structures were made with the conducting rods supported by a dielectric rod; the wavelengths are therefore likely to be smaller than for the corresponding structures as used in aerials, and this means that the theoretical beam width is too large.

(5) GENERAL DISCUSSION

The theory given here for predicting the radiation patterns of end-fire aerials is based on a physically possible method of operation. In the cases where a direct check with experimental results has been possible, good agreement is found and it appears that the method makes possible reasonably accurate estimates of the radiation properties of end-fire aerials. Further, from the postulated radiation mechanism, certain general characteristics of such aerials can be predicted. The theory rests on the assump-

tion that there is an effective radiating area at the end of the rod, whose size determines the beam width. It follows that interaction between dielectric-rod aerials occurs if one aerial is within the effective radiating area of the other. A good indication of the likelihood of interaction can be obtained by the rough rule that dielectric rods interact if they are placed nearer to each other than the distance corresponding to the aperture width of a horn which would produce the same beam width. This rule is quite well known to aerial engineers, but does not appear to have been stated in the literature.

The main topic which requires further discussion is the effect of varying the rod diameter. It is apparent that the dipole mode can be most efficiently launched if the rod diameter is large, but that the beam width of the radiation from the free end is reduced by decreasing the diameter. This suggests the use of a gradually tapered rod as shown in Fig. 8(*a*), which, as mentioned in Section 3, appears to be the direct equivalent of the electromagnetic horn. In practice, however, a gradual tapering of this kind does not lead to a beam width much smaller than that corresponding to the thickest part of the rod. The reason for this is that the dipole wave travelling along the rod is so loosely bound to the guiding structure that the slightest discontinuity, such as the taper, leads to appreciable radiation. The tapered rod in fact behaves as a leaky waveguide and is thus operating in the manner discussed by Mallach.

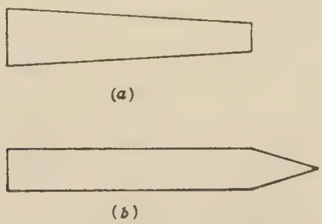

Fig. 8.—Tapered dielectric rods.
(*a*) Gradual taper behaving as a leaky waveguide.
(*b*) Abrupt taper for impedance matching.

Calculations based on Mallach's theory are therefore appropriate to the tapered rod, provided that the rate at which the energy leaks from the rod is reasonably constant along its length. Halliday and Kiely obtain better agreement between the leaky waveguide theory and experiment for tapered rods than for those of uniform diameter. A second common form of tapering is shown in Fig. 8(*b*), where a uniform rod is terminated by an abrupt taper. The bulk of the energy travelling along the uniform section of the rod in the dipole mode will be radiated at the discontinuity between this section and the taper. The radiation pattern is therefore likely to be the same as for a cylindrical rod with a flat end, and this is found in practice. The taper does, however, have a beneficial effect in that it reduces the amplitude of the wave reflected back along the rod towards the feed and thus improves the impedance properties of the aerial.

An interesting extension of end-fire aerials has been described by Simon and Weill,[18] who have developed an aerial of the form shown in Fig. 9. The principle of the aerial is that radiation occurs at each of the discontinuity planes, A, B, etc., where two guiding sections of differing properties join together. The power radiated at each of these planes can be controlled by the magnitude of the discontinuity between the adjacent sections, and the phase differences between successive points of radiation by the length of the intervening guiding section. Simon and Weill have used, as guiding structures, arrays of spaced conducting discs whose diameters are varied at the discontinuity planes. Half-power beam widths down to 7° have been obtained. The aerial functions essentially in the manner described for the Yagi in Section 2.1, the discontinuity planes corresponding to the director elements. The interaction between these planes is, however, much less than that between director elements, allowing much greater freedom in controlling the radiation at each individual radiating point. The interaction in the case of the Yagi elements is, in fact, sufficient to cause the propagation of a surface wave along the structure, as verified by the experiments referred to in Section 4.3.

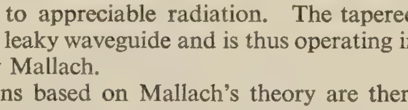

Fig. 9.—Arrangement of end-fire aerial developed by Simon and Weill.
The sections S_1, S_2, etc., are guiding structures of different properties, and radiation occurs at each of the discontinuity planes, A, B, etc.

In conclusion, it should be stressed that the argument presented here depends to a large extent on the value assumed for the amount of energy radiated directly from the feed end of the aerial. The close agreement between the theoretical and experimental patterns of Fig. 7 can be taken as strong evidence that the assumed value of 6% is likely to be nearly correct, for the size of rod considered. Further, this value is of the order predicted by studies of the launching efficiency of other types of surface wave. So far, however, there is no direct evidence as to its magnitude.

(6) ACKNOWLEDGMENT

The work described in the paper was carried out in the Electrical Engineering Department of the Imperial College of Science and Technology, and the authors are indebted to Dr. Willis Jackson for the facilities provided there. During the course of the work, Dr. Spector held a British Council Scholarship and also acknowledges the assistance of the Technion Society of Great Britain for a special grant to complete the work.

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(8) APPENDIX. THE RADIATION OF A PATTERN DIELECTRIC ROD AERIAL

uppose a dielectric rod, of radius a and relative permittivity is operated at a free-space wavelength such that only the ninant mode, the dipole one, can propagate freely along the . If the rod is terminated in the plane $z = 0$, z being the ordinate measured along the rod axis, the field distribution his plane will correspond closely to that of the dipole mode, vided that the mechanism used to launch this mode is ficiently far from the plane $z = 0$. The field strengths fall rapidly with increasing radial distance ρ from the rod axis the field distribution is for all practical purposes restricted a finite area of the plane $z = 0$. The radiated field in the on $z \geqslant 0$ can then be calculated from this distribution using Fourier transform method, which is well established for rtures delineated by a well-defined boundary. The first step uch a calculation is to determine the field distribution for the ole mode.

) The Field Distribution for a Dipole Mode propagating on a Dielectric Rod

tratton[19] has given a general treatment of the properties of ndrical waves such as occur on the dielectric rod, and his ilts can be used to derive general expressions for the com- ents of the electric and magnetic fields. Detailed numerical ulations for the dipole mode have been carried out by ndler[9] and Elsasser,[10] and give values for the wavelength λ_g the mode as a function of the free-space wavelength λ_0 and radius a of the rod. For the present problem, it suffices to centrate on the case when the bulk of the energy travels out- the rod, and then, as may be expected on physical grounds, wavelengths λ_0 and λ_g are nearly equal. The general expres- s for the field components simplify considerably if approxi- ions are made for this condition. Only the transverse ponents of the electric field need be considered, and from results in the above references these can be written:

) Inside the rod, i.e. $\rho \leqslant a$:

$$E_\rho = A[J_1'(k_1\rho) + GJ_1(k_1\rho)/k_1\rho]\cos\phi \quad . \quad . \quad . \quad (3)$$

$$E_\phi = -A[J_1(k_1\rho)/k_1\rho + GJ_1'(k_1\rho)]\sin\phi \quad . \quad . \quad . \quad (4)$$

i) Outside the rod, i.e. $\rho \geqslant a$:

$$E_\rho = -FA[K_1'(k_0\rho) + GK_1(k_0\rho)/k_0\rho]\cos\phi \quad . \quad . \quad (5)$$

$$E_\phi = FA[K_1(k_0\rho)/k_0\rho + GK_1'(k_0\rho)]\sin\phi \quad . \quad . \quad . \quad (6)$$

 these expressions, a common factor $\exp(-j\beta z + j\omega t)$ has been omitted: the prime denotes differentiation with respect to the argument. The various symbols are defined as follows:

A is an arbitrary amplitude constant.

β is the phase-change coefficient for the guided wave along the rod.

$$G = \frac{\beta_0}{\beta}\left[\frac{\epsilon_r J_1'(k_1a)}{k_1J_1(k_1a)} + \frac{K_1'(k_0a)}{k_0K_1(k_0a)}\right]^{1/2} \Bigg/ \left[\frac{J_1'(k_1a)}{k_1J_1(k_1a)} + \frac{K_1'(k_0a)}{k_0K_1(k_0a)}\right]^{1/2} \quad . \quad . \quad . \quad . \quad (7)$$

β_0 is the phase-change coefficient of a plane wave in free space.

$$k_0^2 = \beta^2 - \beta_0^2 \quad . \quad . \quad . \quad . \quad (8)$$

$$k_1^2 = \epsilon_r\beta_0^2 - \beta^2 \quad . \quad . \quad . \quad . \quad (9)$$

and $F = k_1J_1(k_1a)/k_0K_1(k_0a)$ . . . . . (10)

The phase-change coefficient, β, depends on β_0, a and ϵ_r, the relative permittivity of the material of the rod through eqns. (8) and (9), together with[20]

$$\frac{\epsilon_r J_0(k_1a)J_2(k_1a)}{J_1^2(k_1a)} + \frac{(\epsilon_r+1)J_1'(k_1a)}{J_1(k_1a)}\left[\frac{k_1}{k_0^2a} + \frac{k_1K_0(k_0a)}{k_0K_1(k_0a)}\right] - \frac{K_0(k_0a)}{K_1^2(k_0a)}\left[\frac{k_1^2}{k_0^2}K_0(k_0a) + \frac{2k_1^2}{k_0^3a}K_1(k_0a)\right] + \frac{(\epsilon_r+1)}{k_0^2a^2} = 0 \quad (11)$$

In calculating the radiation pattern of the aerial, it is more convenient to work in terms of Cartesian co-ordinates, the field components E_x and E_y being related to E_ρ and E_ϕ by the equations

$$E_x = E_\rho\cos\phi - E_\phi\sin\phi \quad . \quad . \quad . \quad . \quad (12)$$

$$E_y = E_\rho\sin\phi + E_\phi\cos\phi \quad . \quad . \quad . \quad . \quad (13)$$

The directions of the x- and y-axes are shown in Fig. 3. Substitution from eqns. (2) to (6) gives

$$\rho \leqslant a: \quad E_x = A[J_1'(k_1\rho)(\cos^2\phi + G\sin^2\phi) + J_1(k_1\rho)(G\cos^2\phi + \sin^2\phi)/k_1\rho] \quad . \quad (14)$$

$$E_y = A[J_1'(k_1\rho) - J_1(k_1\rho)/k_1\rho](1 - G)\sin\phi\cos\phi \quad (15)$$

$$\rho \geqslant a: \quad E_x = -FA[K_1'(k_0\rho](\cos^2\phi + G\sin^2\phi) + K_1(k_0\rho)(G\cos^2\phi + \sin^2\phi)/k_1\rho] \quad . \quad (16)$$

$$E_y = -FA[K'(k_0\rho) - K_1(k_0\rho)/k_0\rho](1 - G)\sin\phi\cos\phi \quad . \quad (17)$$

The expressions in the above equations are exact. Approximations will now be made on the assumption that the phase-change coefficient β of the wave on the rod is almost equal to the free-space phase-change coefficient, β_0, so that k_0 can be taken as a small quantity. The modified Bessel function $K_1(k_0a)$ is then almost equal to $1/k_0a$, so that eqn. (7) gives

$$G \simeq \frac{\beta_0}{\beta}\left[1 - \frac{(\epsilon_r - 1)J_1'(k_1a)k_0^2a}{2k_1J_1(k_1a)}\right] \quad . \quad . \quad . \quad (18)$$

Also, from eqn. (8),

$$\beta \simeq \beta_0[1 + (k_0^2/2\beta_0^2)] \quad . \quad . \quad . \quad . \quad (19)$$

To the same order of approximation, eqn. (11) becomes

$$\frac{(\epsilon_r+1)J_1'(k_1a)}{J_1(k_1a)} = 2k_1aK_0(k_0a) - \frac{(\epsilon_r+1)}{k_1a} \quad . \quad (20)$$

which can be substituted in eqn. (18) to give

$$G \simeq 1 - \frac{k_0^2}{2\beta_0^2} - \frac{(\epsilon_r-1)k_0^2a^2}{2(\epsilon_r+1)}\left[2K_0(k_0a) - \frac{(\epsilon_r+1)}{(k_1a)^2}\right]$$

$$= 1 - \frac{\epsilon_r - 1}{\epsilon_r + 1}K_0(k_0a)(k_0a)^2 \quad . \quad . \quad . \quad . \quad . \quad (21)$$

The function $K_0(k_0a)$ tends to $\log_\varepsilon(1/k_0a)$ for small values of k_0a so that G tends to unity. G is larger than 0·9, provided that k_0a is less than 0·5 and ϵ_r has the value 2·56 appropriate to polystyrene. The corresponding ratio of the rod diameter to the free-space wavelength is 0·4.

For the remainder of the calculations, G is taken as unity. Although this value is in error at the largest rod diameters considered, the resulting error in the calculated radiation patterns can be shown by a fuller analysis to be relatively less severe. If G is replaced by unity in eqns. (14)–(17), and at the same time use is made of the recurrence relations

$$J_1'(k_1\rho) + \frac{J_1(k_1\rho)}{k_1\rho} = J_0(k_1\rho) \quad . \quad . \quad . \quad . \quad (22)$$

$$K_1'(k_0\rho) + \frac{K_1(k_0\rho)}{k_0\rho} = -K_0(k_0\rho) \quad . \quad . \quad . \quad (23)$$

there results

$$E_x = AJ_0(k_1\rho) \qquad \text{for } \rho \leqslant a \quad . \quad . \quad . \quad . \quad . \quad (24)$$

$$= FAK_0(k_0\rho) \qquad \text{for } \rho \geqslant a \quad . \quad . \quad . \quad . \quad . \quad (25)$$

$$E_y = 0 \qquad \text{for all } \rho \quad . \quad . \quad . \quad . \quad . \quad (26)$$

(8.2) Calculation of the Radiation Pattern

The radiation pattern of a field distribution specified by the function $F(\rho, \phi)$ in the aperture plane is given by[4]

$$g(\theta, X) = (1 + \cos\theta)\int_0^{2\pi} d\phi \int_0^\infty d\rho\rho F(\rho, \phi)\exp[j\beta_0\rho\sin\theta\cos(X - \phi)]$$

where $g(\theta, X)$ is the field intensity at a point on the surface of a sphere specified by the polar angles θ and X. If, as in the present case, $F(\rho, \phi)$ is independent of ϕ, the integration with respect to ϕ can be performed immediately, giving

$$g(\theta, X) = 2\pi(1 + \cos\theta)\int_0^\infty \rho F(\rho)J_0(\beta_0\rho\sin\theta)d\rho \quad . \quad (27)$$

which shows that the radiation pattern is independent of X.

The function $F(\rho)$ is defined by eqns. (24) and (25), so that

$$g(\theta) = 2\pi A(1 + \cos\theta)\left[\int_0^a \rho J_0(k_1\rho)J_0(\beta_0\rho\sin\theta)d\rho + F\int_a^\infty \rho K_0(k_0\rho)J_0(\beta_0\rho\sin\theta)d\rho\right] \quad . \quad (28)$$

Each integral is of the Lommel type, and integration gives

$$g(\theta) = 2\pi Aa^2(1 + \cos\theta)\left[\frac{k_1aJ_1(k_1a)J_0(u) - uJ_1(u)J_0(k_1a)}{k_1^2a^2 + u^2} + k_1J_1(k_1a)\frac{k_0aK_1(k_0a)J_0(u) - uK_0(k_0a)J_1(u)}{k_0K_1(k_0a)(k_0^2a^2 + u^2)}\right] \quad . \quad (29)$$

where $$u = \beta_0 a\sin\theta \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (30)$$

The expression in eqn. (29) can be readily evaluated with the help of tables.

(8.3) Note on the Approximation made in Section 8.1

The results given in eqns. (24)–(26) are adequate for the calculation of the radiation pattern, but it is evident that they cannot satisfy the boundary conditions at the surface of the rod. The reason for this breakdown of the approximation is that, when ρ is nearly equal to a, the terms $K_1(k_0\rho)/(k_0\rho)$ and $K_1'(k_0\rho)$ are of the same order as $(1 - G)$. If allowance is made for this, it is found that there is an additional field outside the rod, with both x- and y-components proportional to $(a/\rho)^2$. This field varies with ϕ in the same way as the field caused by a distribution of dipoles along the rod axis, the dipoles being directed parallel to the x-axis. The rapid decay of the field with ρ makes it insignificant as far as the radiation pattern calculated for the polarization in the x-direction is concerned. There will, however, be cross-polarized radiation, with a null in the forward direction.

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A NON-RESONANT WAVEGUIDE WINDOW

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SUMMARY

From simple waveguide theory it is shown that the characteristic ve impedance of air-filled (or evacuated) and dielectric-filled wave- ides of equal cross-section propagating E-modes can be matched any free-space wavelength and relative permittivity of the dielectric. r low-permittivity dielectrics, reflectionless transmission may be tained for a bandwidth of 20%. It is possible to design a waveguide ndow on these lines, the thickness of which—unlike that of the onant-iris window—is not critical. The simplicity of the structure pears suitable for high-power microwave valves. A universal design- art is derived from which the window diameter for the matched ndition may be read off directly for any free-space wavelength and ative permittivity of the dielectric. A window design with broad- nd mode-transformers is described.

LIST OF PRINCIPAL SYMBOLS

Z_{0a} = Characteristic wave impedance of air-filled (or evacuated) waveguide.
Z_{0d} = Characteristic wave impedance of dielectric-filled waveguide.
ϵ_d = Relative permittivity of dielectric.
λ_0 = Free-space wavelength at frequency f_0.
λ_c = Cut-off wavelength of waveguide.
λ_{0m} = Free-space wavelength for matched condition.
λ_{gd} = Guide wavelength in dielectric-filled waveguide for free-space wavelength λ_0.
λ_{gdm} = Guide-wavelength in dielectric-filled waveguide for matched free-space wavelength λ_{0m}.
β_d = Propagation coefficient of dielectric-filled waveguide.
l = Thickness of dielectric-filled section.

(1) INTRODUCTION

n modern high-power microwave oscillators and amplifiers power is usually extracted directly from a resonant cavity d the full-size waveguide is continued inside the valve. The ed then arises for a vacuum-tight joint to complete the valve elope, which is obtained by sealing a low-loss dielectric ndow across the inside of the waveguide. In order to obtain ectionless transmission, the anti-resonant properties of the veguide iris may be employed, and for some time this has been most commonly used coupling device. While satisfactory most purposes, however, it does suffer from several dis- antages: in particular, it is required to function as both a uum seal and a lossless and reflectionless waveguide element, l in many cases the mechanical design is dictated by electrical siderations. Although there is a simple relation between iris meter and glass-window thickness,[1] it has been shown in ctice[2] that, if flared waveguide sections are used to reduce the tage gradient at the window for high peak-power transmission l urther complications are introduced by quarter-wave chokes, re 's only one possible pair of dimensions for a given waveguide . As a result, particularly at very high frequencies, the window thickness must be kept within extremely close limits. If ceramics are used, their high permittivities necessitate even more complicated design procedures.[3]

However, recent experience in the development of high-power microwave valves has shown the desirability of much higher baking temperatures than are customary at present. In this respect, ceramics offer more attractive possibilities than the sealing glasses in common use and ceramic-to-metal vacuum seals are receiving increasing attention.

If the mechanical and electrical functions of the window could be separated, the valve engineer might optimize the mechanical design of the vacuum seal. The device described in the paper[4] goes some way to achieving this, since the window thickness is now no longer subject to close tolerances. Instead of relying on the anti-resonant properties of the iris, reflectionless transmission is obtained by the correct choice of waveguide dimensions to ensure perfect matching at the operating frequency between air-filled (or evacuated) and dielectric-filled sections. Although matching is restricted to E-mode operation, necessitating the use of mode transformers for most applications, this may well be compensated by the extreme simplicity of the window geometry.

(2) MATCHING PARAMETERS

For a waveguide of uniform cross-section filled with a dielectric assumed to be lossless, the characteristic wave impedance is given by[5]

$$Z_{0d} = 377/\sqrt{[\epsilon_d - (\lambda_0/\lambda_c)^2]} \quad . \quad . \quad . \quad . \quad (1)$$

for H-waves

and

$$Z_{0d} = 377\sqrt{[\epsilon_d - (\lambda_0/\lambda_c)^2]}/\epsilon_d \quad . \quad . \quad . \quad . \quad (2)$$

for E-waves

If Z_{0d} is plotted against λ_0/λ_c for different values of ϵ_d, families of curves are obtained as shown by Barlow and Cullen,[6] where any two curves intersect for E-modes and diverge for H-modes. It follows that the characteristic wave impedance of two waveguides of equal cross-section, propagating the same E-wave but filled with different dielectrics, can be made equal at a value of λ_{0m}/λ_c which depends only on ϵ_d.

For the particular case of an air-filled (or evacuated) and a dielectric-filled waveguide, this condition holds when

$$377\sqrt{[1 - (\lambda_{0m}/\lambda_c)^2]} = 377\sqrt{[\epsilon_d - (\lambda_{0m}/\lambda_c)^2]}/\epsilon_d \quad . \quad (3)$$

from which $$\lambda_{0m}/\lambda_c = \sqrt{[\epsilon_d(1 - \epsilon_d)/(1 - \epsilon_d^2)]} \quad . \quad . \quad . \quad (4)$$

It is obvious that, when $\epsilon_d = 1$, no problem of matching arises and λ_{0m}/λ_c is indeterminate. For $\epsilon_d > 1$, eqn. (4) reduces to

$$\lambda_{0m}/\lambda_c = \sqrt{[\epsilon_d/(1 + \epsilon_d)]} \quad . \quad . \quad . \quad . \quad (5)$$

This relation was derived by Lamont[7] and used by Pincherle,[8] who investigated the effects of dielectrics in waveguides and obtained theoretical curves of the reflection coefficient as a function of frequency for junctions of air-filled and infinitely long dielectric-filled waveguides. It has been used recently in the design of dielectric-loaded linear accelerators.[9]

For a given E-mode, relative permittivity and free-space

itten contributions on papers published without being read at meetings are ed for consideration with a view to publication.

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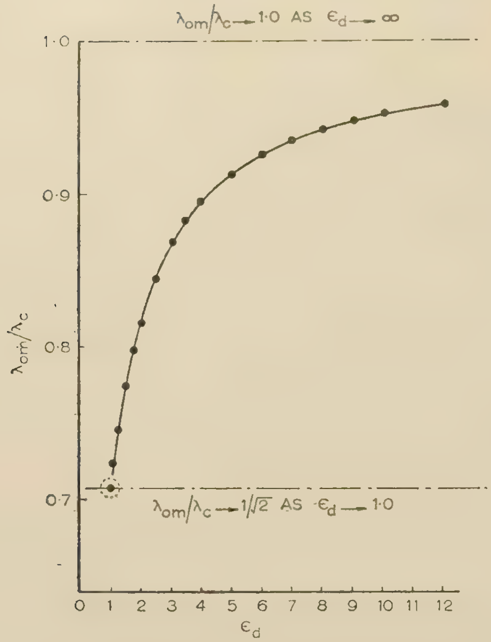

Fig. 1.—Dependence of λ_{0m}/λ_c on ϵ_d.

wavelength, the guide dimensions may be calculated from eqn. (5); λ_{0m}/λ_c is plotted as a function of ϵ_d in Fig. 1. Provided that this relation is satisfied, a dielectric slab may be inserted in the waveguide, irrespective of thickness, without affecting the characteristic wave impedance.

(3) BANDWIDTH

Although perfect matching is obtained at the free-space wavelength λ_{0m}, it is of interest to establish the range of wavelength for which a given window introduces a voltage standing-wave ratio (v.s.w.r.) of 1·2 or less.

Case (a).—Air-filled (or evacuated) waveguide terminated by infinitely long loss-free dielectric-filled guide.

The input impedance of this arrangement is purely resistive. Therefore, for a v.s.w.r. $\leqslant$ 1·2,

$$Z_{0a}/Z_{0d} \leqslant 1\cdot2 \geqslant Z_{0d}/Z_{0a} \quad . \quad . \quad . \quad . \quad (6)$$

But, from eqn. (2),

$$(Z_{0a}/Z_{0d})^2 = \epsilon_d^2[1 - (\lambda_0/\lambda_c)^2]/[\epsilon_d - (\lambda_0/\lambda_c)^2] \quad . \quad (7)$$

whence $$\lambda_0/\lambda_c = \epsilon_d[\epsilon_d - (Z_{0a}/Z_{0d})^2]/[\epsilon_d^2 - (Z_{0a}/Z_{0d})^2] \quad . \quad (8)$$

The required limits of λ_0/λ_c are found by putting Z_{0a}/Z_{0d} equal to 1·2 and 1/1·2 respectively. The bandwidth as calculated from eqn. (8), plotted as a function of ϵ_d in Fig. 2, increases rapidly as ϵ_d decreases, being infinite for $\epsilon_d = 1\cdot2$. For $\epsilon_d < 1\cdot2$, the v.s.w.r. is always less than 1·2 [from eqn. (7)].

Case (b).—Air-filled (or evacuated) waveguide containing dielectric slab of finite thickness.

The bandwidth of this configuration can be obtained from a simple transmission-line analogy. Consider an infinitely long air-filled waveguide containing a lossless dielectric slab of thickness l.

The input impedance at the air-dielectric interface is given by

$$Z = Z_{0d}\frac{Z_{0a}\cos\beta_d l + jZ_{0d}\sin\beta_d l}{Z_{0d}\cos\beta_d l + jZ_{0a}\sin\beta_d l} \quad . \quad . \quad . \quad (9)$$

where Z_{0a} and Z_{0d} may be calculated from eqn. (2).

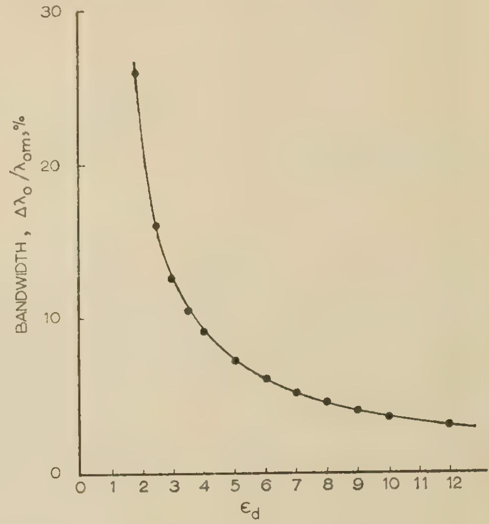

Fig. 2.—Bandwidth of air-filled guide terminated by infinitely long loss-free dielectric-filled guide at free-space wavelength λ_{0m}.

Let $$l = \rho\lambda_{gdm} \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (10)$$

where ρ is a constant.

Then $$\beta_d = 2\pi/\lambda_{gd}, \text{ and } \beta_d l = 2\pi\rho\lambda_{gdm}/\lambda_{gd} \quad . \quad . \quad (11)$$

But,[10] $$\lambda_{gdm} = \lambda_{0m}/\sqrt{[\epsilon_d - (\lambda_{0m}/\lambda_c)^2]} \quad . \quad . \quad . \quad (12)$$

and $$\lambda_{gd} = \lambda_0/\sqrt{[\epsilon_d - (\lambda_0/\lambda_c)^2]} \quad . \quad . \quad . \quad . \quad (13)$$

Hence

$$\beta_d l = 2\pi\rho\sqrt{\left\{\frac{[\epsilon_d - (\lambda_0/\lambda_c)^2]}{[\epsilon_d - (\lambda_{0m}/\lambda_c)^2]}\right\}}\frac{\lambda_{0m}}{\lambda_0}$$

$$= 2\pi\rho\sqrt{\left\{\frac{[\epsilon_d - (\lambda_{0m}/\lambda_c)^2(\lambda_0/\lambda_{0m})^2]}{[\epsilon_d - (\lambda_{0m}/\lambda_c)^2]}\right\}}\frac{1}{\lambda_0/\lambda_{0m}} \quad . \quad (14)$$

By substituting eqn. (14) in eqn. (9), the reflection coefficient and v.s.w.r. as a function of λ_0/λ_{0m} can be computed, and from this the range of wavelength for which the v.s.w.r. is less than 1·2 be obtained.

Since this procedure is somewhat laborious, only special cases of practical interest will be considered. Thus, the most unfavourable conditions are met when a dielectric of thickness $\lambda_{gdm}/4$ is used, where reflections from both air-dielectric interfaces are in phase. The v.s.w.r. introduced by such a window of quartz or Corning 7070 glass ($\epsilon_d = 4$) is plotted against λ_0/λ_{0m} in Fig. 3(*a*), where a bandwidth of over 4% is obtained. In practice, very much thinner windows are used for high-power applications, ranging in thickness from 0·02 to 0·04λ_0.[11] The v.s.w.r. for an alumina window of thickness 0·04λ_0 is plotted against λ_0/λ_{0m} for comparison in Fig. 3(*b*), giving a bandwidth of 4·2%. In general, the bandwidth increases with decreasing window thickness, and for a given thickness increases as ϵ_d is reduced.

(4) EXPERIMENTAL EQUIPMENT

Some confirmation of the theoretical results outlined above was obtained with discs of polystyrene cemented in a circular brass waveguide propagating the E_{01}-mode, which was chosen for practical reasons. For convenience, measurements were made in the 3 cm band. The E_{01}-mode was excited by means of

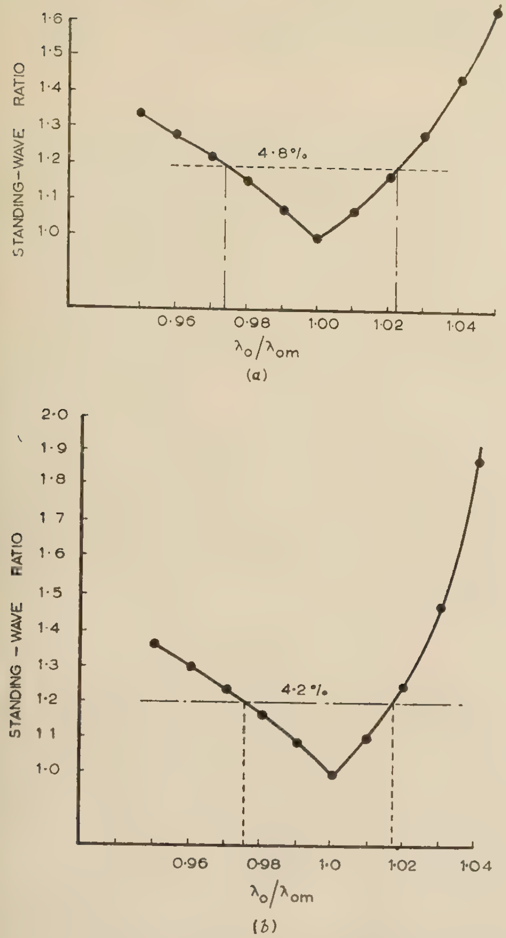

Fig. 3.—Standing-wave ratios of dielectric windows.

(a) $\epsilon_d = 4$, $l = 0 \cdot 25\lambda_{gdm}$.
(b) $\epsilon_d = 9$, $l = 0 \cdot 04\lambda_0 = 0 \cdot 114\lambda_{gdm}$.

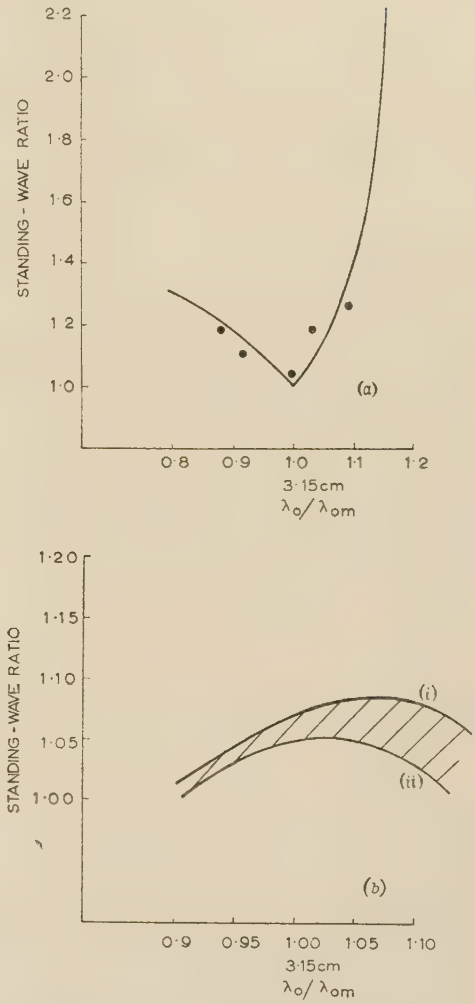

Fig. 4.—Standing-wave ratios of polystyrene windows ($\epsilon_d = 2 \cdot 54$).

(a) $l = 0 \cdot 25\lambda_{gdm} = 0 \cdot 23$ in.
— Calculated. ● Measured.
(bi) $l = 0 \cdot 5\lambda_{gdm} = 0 \cdot 46$ in. (bii) $l = 0 \cdot 07\lambda_{gdm} = 0 \cdot 065$ in.

›axial probe, and the v.s.w.r. introduced by the dielectric low was measured with a probe travelling in an axially ›d slot. A conical wood load provided a reflectionless ination. Owing to the difficulty of maintaining a uniform ılar cross-section over the whole length of the milled slot, accuracy of this type of standing-wave indicator was not cient to obtain reliable measurements of standing-wave ratios ›r than 1·05. It was, however, considered satisfactory to onstrate the overall performance of the polystyrene windows d.

(5) EXPERIMENTAL RESULTS

ıe experimental and theoretical curves for a polystyrene ow ($\epsilon_d = 2 \cdot 54$) of thickness $\lambda_{gdm}/4$ are shown in Fig. 4(a). ew of the limited accuracy of the measuring equipment, the ement obtained is good. As stated in Section 3, this repre- the most unfavourable condition. For comparison, the ormances of two windows of thickness $0 \cdot 5$ and $0 \cdot 07\lambda_{gdm}$ ctively are shown in Fig. 4(b). In both cases the v.s.w.r. was han 1·1 for the whole frequency range covered by the local ator (3–3·3 cm), but owing to the limitations of the ing-wave indicator mentioned above, there was considerable r of the measured values, the reflection from the thinner ow being generally less than that from the thicker. The conclusion arising from these results is that the bandwidth

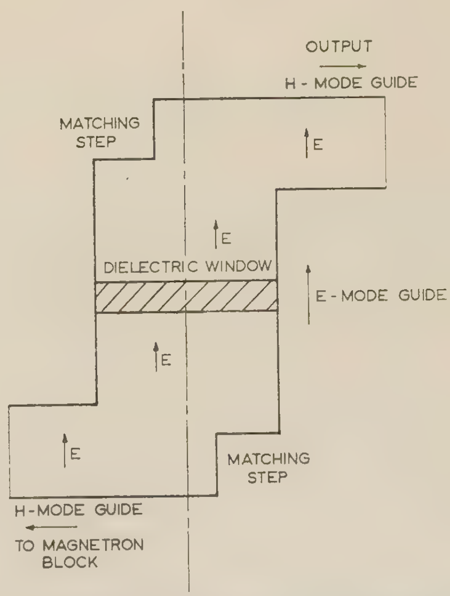

Fig. 5.—Matched window with mode-transformers.

is considerably greater than that covered by the experiment, and unless extremely broad-band performance is desired, the window thickness is not critical. In view of the simplicity of the arrangement, it should be possible to design a satisfactory dielectric-to-metal vacuum seal on these lines; however, no facilities were available to investigate techniques of this nature.

(6) A PRACTICAL MATCHED WAVEGUIDE WINDOW

Since matching can be obtained for E-modes only, it will be necessary for most practical purposes to introduce a transition from the H_{01}-mode in a rectangular waveguide. However, the properties of H_{01}/E_{01} mode-transformers have been studied extensively, and practical broad-band high-power designs have been developed.[12] For this reason, the E_{01}-mode was chosen for the experiments described previously. Following these considerations, a broad-band output system is suggested where the window is sealed in the circular section terminated by mode transformers as shown in Fig. 5. This arrangement would also prevent damage to the window by stray electrons or X-rays.[13]

As a practical example, the design of an alumina window ($\epsilon_d = 9$) matched at the centre wavelength of 3 cm will be given.

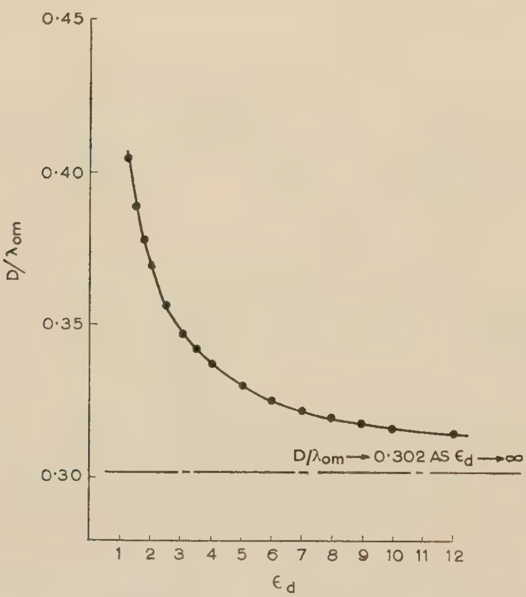

Fig. 6.—Universal design chart: waveguide diameter D (in) as a function of λ_{0m} (cm) and ϵ_d.

From Fig. 1, for $\epsilon_d = 9$, $\lambda_{0m}/\lambda_c = 0{\cdot}948$, i.e. $\lambda_c = 3{\cdot}0/0{\cdot}948$ cm. But for the E_{01}-mode, $\lambda_c = 2{\cdot}61D/2$, where D is the guide diameter. Therefore $D = 3{\cdot}0 \times 2/0{\cdot}948 \times 2{\cdot}61 = 2{\cdot}43$ cm. A universal design chart has been developed on these lines for the matched condition, where as a matter of convenience the ordinate represents the ratio between the waveguide diameter in inches and the matched wavelength in centimetres. This function [is] plotted against ϵ_d in Fig. 6. Thus the waveguide diameter ca[n] be determined for any relative permittivity and free-spa[ce] wavelength.

(7) CONCLUSIONS

For most purposes the performance of the commonly use[d] resonant-iris window is perfectly adequate, but where extrem[e] requirements make it difficult to use conventional structures an[d] materials the design suggested in the paper may offer a possib[le] alternative. It has been shown conclusively that, for a give[n] bandwidth, mechanical tolerances are far less critical than tho[se] of other output systems, and, for this reason alone, the devi[ce] may merit further practical development.

(8) ACKNOWLEDGMENT

Thanks are due to Mr. L. C. Robinson, who drew attention [to] the problem, and to Dr. G. B. Walker for helpful discussion[s]. The authors wish to acknowledge the award of a Fellowship b[y] the Atomic Energy Research Establishment (A. E. B.) and of [a] Maintenance Grant from the Department of Scientific an[d] Industrial Research (J. T. H.).

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Paper No. 2239 R
Jan. 1957

...DING OF LONG-DISTANCE RADIO SIGNALS AND A COMPARISON OF SPACE- AND POLARIZATION-DIVERSITY RECEPTION IN THE 6–18 Mc/s RANGE

By G. L. GRISDALE, Ph.D., J. G. MORRIS, B.Sc., and D. S. PALMER, M.A.

(*The paper was first received* 18*th July, and in revised form* 11*th September*, 1956.)

SUMMARY

...he rapid component of the variation of signal strength from ...ant radio transmitters operating in the 6–18 Mc/s frequency band ...been examined experimentally, and the results of the measurements ...compared with various statistical laws which have been suggested ...n theoretical considerations. The rapid component of fading ...ound to agree closely with a Rayleigh distribution. A charac...tic time-constant for the rapid fading is suggested and its value ...rmined for a number of distant stations.

...leasurements have also been made of the correlation between ...al strength variations in two spaced aerials and in two aerials at ...same place but set at right angles. From these results a comparison ...pace and polarization diversity is made. It is concluded that both ...ems could, on average, give equal performance. The two diversity ...ems were also compared in operation on a telegraph circuit between ...ados and England, the number of telegraphic distortions being ...ted in each aerial in turn. There was no significant difference ...een the two systems.

...y combining the fading-law results and the diversity-correlation ...ts, the improvement due to diversity may be expressed in terms ...n equivalent power gain; the gain is higher on communication ...uits which are initially good than on those which are poor.

LIST OF SYMBOLS

A, B, C = Measured fractions of time.
d = Distance between aerials, m.
G = Power gain, dB.
h = Signal amplitude expressed as a fraction of the most probable amplitude.
I_0 = Bessel function.
K = Index defined in text.
m = Constant.
N = Half the number of excursions of signal amplitude across a datum level.
n = Number of channels in a diversity receiving system.
P = Proportion or probability.
p = Number of pulses.
R, ρ = Correlation or auto-correlation coefficient.
..., S'', S_1, S_2 = Instantaneous values of signal amplitude.
...$_m$, S_{m1}, S_{m2} = Median values of signal amplitude.
T, t, τ = Time.
X = Constant, with dimension of distance.
x, y = Signal amplitudes defined in text.
θ = Proportion of time that a signal is unusable.
σ = Fading speed defined in text.

(1) INTRODUCTION

...rect radio communication over long distances uses the ...ency range 3–25 Mc/s, the radio wave being propagated ...d the curvature of the earth by the ionospheric layers. Owing to the irregular and changing nature of these layers and interference between waves travelling along different paths between transmitter and receiver, the received signal amplitude fluctuates or fades. Some oscillograph records of the rapid fading of a single-frequency signal are shown in Fig. 1, the total

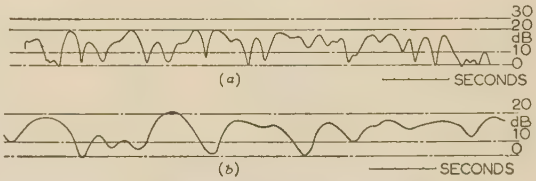

Fig. 1.—Fading of single-frequency transmissions.

(*a*) Shepparton, Australia, 27th October, 1954. 1410 G.M.T. 11 900 kc/s.
(*b*) Sackville, Canada, 7th October, 1954. 1358 G.M.T. 15 090 kc/s.

excursion of 20–30 dB being covered in a few seconds. In addition to the rapid variations mainly caused by wave interference, slower changes, not examined in the paper, are caused by fluctuations in the attenuation of the wave paths between the transmitter and receiver.

The degradation of communication caused by rapid fading is reduced by using diversity reception, i.e. by using two reception channels which are so arranged that the signal does not always fade in both channels together; the channel giving the better signal is automatically selected to provide the signal used for communication.

There are at least four forms of diversity reception,[1] including time-diversity, when the signal is sent through the same channel more than once, and frequency-diversity, when the signal is transmitted and received simultaneously on two frequencies. Though both these systems are in use, neither finds favour, because time-diversity reduces the traffic capacity of the circuit and frequency-diversity increases the frequency band necessary to transmit a given amount of information and probably leads to more transmitter power and complication.

The most-used system is spaced-aerial diversity reception, in which two similar aerials are arranged some wavelengths apart; it is found that the fading of the same signal is not coincident in the two aerials, which can therefore be used for diversity reception.

An alternative arrangement uses independent receiving aerials close together, but differing in direction, so that they respond to different polarization components of the wave. If the fading is in part due to changes of polarization, the mutually inclined aerials will be effective in a diversity system. This arrangement is known as polarization diversity.

By using two receivers with the same phase and gain characteristics, one having a vertical aerial and one a horizontal aerial, it is possible to display on an oscillograph the polarization of a wave arriving at the aerials. Photographs of such polarization diagrams are shown in Fig. 2; the polarization is elliptical and changes rapidly with time. It also changes for a very small fre-

...ten contributions on papers published without being read at meetings are ...d for consideration with a view to publication.
...Grisdale and Mr. Palmer are and Mr. Morris was at the Research Laboratories ...coni's Wireless Telegraph Co., Ltd. Mr. Morris is now at the Government ...unications Headquarters, Cheltenham.

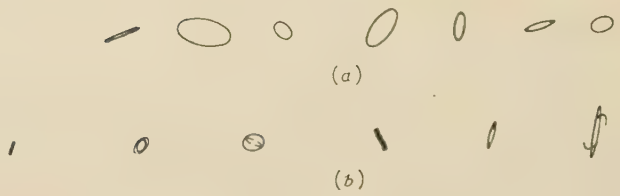

Fig. 2.—Oscillograph records showing polarization changes with time and frequency.

(a) Sackville, Canada. 4th November, 1954. 1327 G.M.T. 15 090 kc/s. (Carrier only.)
(b) Istanbul, Turkey. (T.A.G.) 7th October, 1954. 1447 G.M.T.
Exposures made at 5 sec intervals.

quency shift, as shown by the ellipses corresponding to the mark and space frequencies of a frequency-shift telegraph transmission.

(2) THE FADING LAW

The fading law defines the relation between the proportion of time P that the signal produced in an aerial falls below a level S, and the value of S; statistically it is known as the probability integral. Throughout the following discussion the signal will be assumed to consist essentially of a single frequency, and in the experimental work the bandwidth was so restricted that the single frequency component of the signal was measured.

Van der Pol[3] has considered a signal which is the sum of a large number of components of random phase having the same frequency, and which would have a Rayleigh type of fading law given by

$$P(S) = 1 - \exp\left[-0{\cdot}693\,(S/S_m)^2\right] \quad . \quad . \quad . \quad (1)$$

where $P(S)$ is the probability that the signal will be less than S, and S_m is the median value of the signal. This law is represented by the curve marked 'single channel' in Fig. 3.

The power law suggested by Jelonek, Fitch and Chalk[2] is

$$P(S) \propto S^m \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (2)$$

where m is a constant.

For the low values of $P(S)$ with which we are concerned in communication, the Rayleigh law closely approaches a square law ($m = 2$). Measurements[2] have shown that for a distance of 116 km and a frequency of 1 025 kc/s, the index m was 2, whilst for microwave link experiments a value of 2·6 is quoted. Van Wambeck and Ross[4] have published measurements for frequencies between 7 and 16 Mc/s over a 1 440 km path which suggest a mean index of about 2.

In order to obtain more information about the form of the fading curve for high-frequency transmission over distances of 2 250–16 000 km, a series of measurements was made using the carriers of amplitude-modulated transmissions. The apparatus, which will be described later, was arranged to record the integrated time intervals that the carrier spent below each of four chosen levels during a test period of about 10 min. Over such a period, the distribution is essentially that of the short-period variations; the long-period changes, due to slower variations of the ionosphere, are not reduced by spaced-aerial and polarization diversity systems, and are thus of less interest in this study.

The four results obtained from each of these tests were plotted to show the proportion of time faded as a function of the signal level referred to the median value, so that all the curves would pass through the point $P(S) = 0{\cdot}5$, $S = 1$ (or 0 dB). The points are shown as crosses in Fig. 11, with the Rayleigh distribution drawn as the full line. Down to $P(S) = 0{\cdot}1$, the experimental points are grouped about the Rayleigh curve, but when $P(S) < 0{\cdot}05$ the points indicate less fading than would be given by the Rayleigh law but more than the lognormal curve drawn

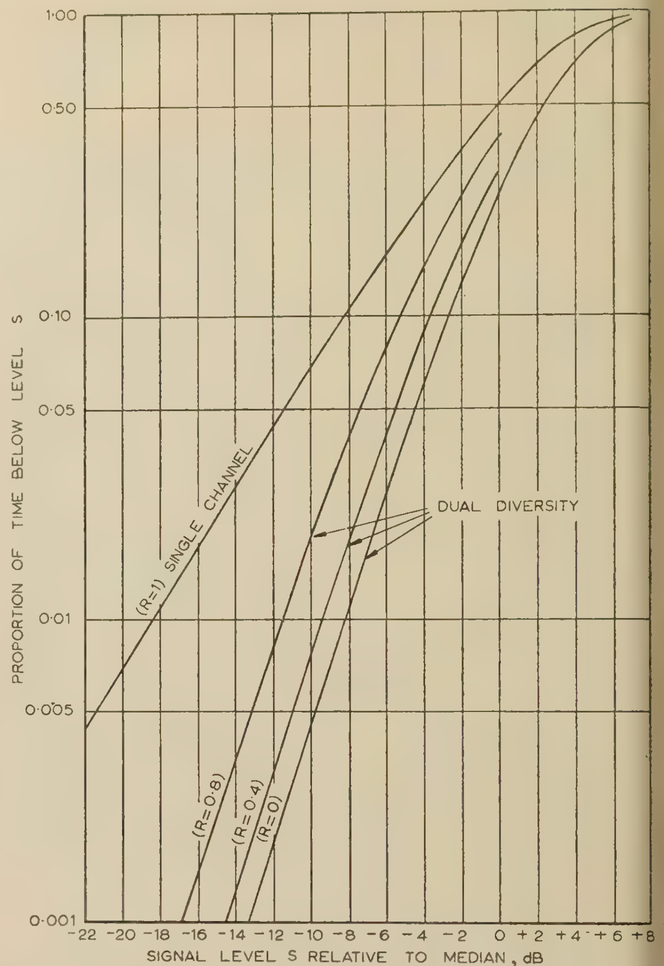

Fig. 3.—Fading characteristics of single-channel and dual-diversity systems with Rayleigh-type fading for different values of correlation coefficient R.

to give the best fit to the experimental points. Results reported to the Comité Consultatif International Radio[5] indicated that for tests covering periods longer than 15 min a lognormal distribution is in fact approached.

Measurements to determine the fading law were made with most of the transmissions used in the associated diversity investigation, and are listed in the Tables. Insufficient data are available to study the variation of the fading law with distance, frequency and hour of measurement; from the small scatter of the observed points this variation is thought not to be great.

At low levels of signal the scatter of the points is greater, but there is no doubt as to the general trend. The scatter is due to the small number of very deep fades; the integration is by a sampling method at a rate of 375 per min, the number of occasions when the signal level is below the set level being recorded. For a 10-min run and a proportional time faded of 0·002, the average count is only 7–8, and this small number could be expected to produce scatter, without, however, introducing any bias in the general trend of the points.

All the possible causes for a departure from the Rayleigh distribution at low levels indicate that less fading than the Rayleigh distribution should be observed, in agreement with the experimental evidence. Thus the presence of sidebands of the transmission which would fade independently of the carrier and which were not attenuated sufficiently by the receiver

ctivity would reduce the fraction of time recorded below as en level. The same would apply for an interfering signal, noise, although this is unlikely to be significant as the transsions were chosen to give clear signals of good strength.

rom the data obtained it was concluded that the Rayleigh ribution might be assumed in making diversity computations.

(3) TIME STRUCTURE OF THE FADING

esides knowing the fading law, the time scale of the fading racteristic is of importance, both for the design of a.g.c. ems and for diversity path-selection apparatus. It can be n from Fig. 1 that the fading is more rapid for the Australian n for the Canadian transmissions. A statistical parameter σ, ed the fading speed, can be defined to describe this property the received signal.

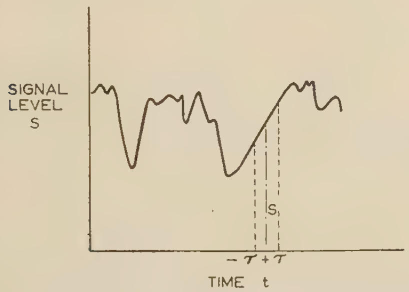

Fig. 4.—Signal strength as a function of time.

ig. 4 is typical of the variation of signal intensity with time: ime t the signal level is S, and at times $t - \tau$ and $t + \tau$ the al intensity is slightly different. The correlation coefficient be calculated for the pairs of values of S at instants t and τ for increasing values of τ; it is called, in this special case, auto-correlation of the curve[6, 7] and is defined by

$$\rho(\tau) = \frac{\overline{S(t)S(t+\tau)} - [\overline{S(t)}]^2}{[\overline{S(t)}]^2 - [\overline{S(t)}]^2} \quad . \quad . \quad . \quad . \quad (3)$$

he auto-correlation measures the average correlation existing ween the signal intensities at two instants apart. The resulting ph of $\rho(\tau)$ as a function of τ is called the auto-correlation ction or correlogram and is characteristic of the fading-time cture. Obviously for $\tau = 0$ the auto-correlation is unity, after a sufficiently long time it is reasonable to suppose that value of the signal is independent of its value at $\tau = 0$,

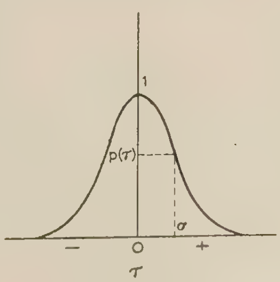

Fig. 5.—Auto-correlation of signal-strength curve.

ρ becomes zero. A smooth curve such as that of Fig. 5 is ained, which may be assumed symmetrical about $\tau = 0$.

or fading signals of Rayleigh distribution it is assumed that can be produced as the r.m.s. of two equal independent ssian distributions each with a Gaussian correlogram; this has some experimental verification.[8] The Gaussian autocorrelogram may be written:

$$\rho(\tau) = \varepsilon^{-\tau^2/2\sigma^2} \quad . \quad . \quad . \quad . \quad . \quad . \quad (4)$$

the constant σ being the fading speed, i.e. the time interval after which the auto-correlation ρ falls to the value $\varepsilon^{-1/2} = 0 \cdot 61$.

The fading speed can be deduced from the experimental measurements of the number of fades, N, occurring in a given time interval, T, since, for a Rayleigh distribution of signal amplitude

$$\sigma = \frac{h(1 - A)}{\sqrt{(2\pi)}} \frac{T}{N} \quad . \quad . \quad . \quad . \quad . \quad . \quad (5)$$

where h is the signal amplitude expressed in terms of the most-probable amplitude, A is the measured fraction of time during which the signal intensity lies below h, and N is half the number of times the signal crosses the datum signal intensity.

The fading speed, σ, is independent of the signal intensity, and, for the probability integral measurements, in which the signal was sampled at four levels, σ should be constant for each recording.

(3.1) Experimental Results

Fig. 6 shows histograms of σ, together with the mean values for each station at each frequency; the standard deviations of the distributions are also indicated. In Fig. 7 the mean fading speed for all the tests for a given station and frequency is written against a point whose co-ordinates are distance and frequency. The stations fall into the same groups according to distance as were found for correlation of fading, discussed later.

(*a*) *European and Transatlantic Transmissions.*

With the exception of the three $9\frac{1}{2}$ Mc/s tests from Sackville (ringed), the mean fading speed is low and does not vary greatly with frequency in the 6–15 Mc/s range.

The mean σ for all these stations is 2·0 sec with a standard deviation of 0·9 if the 15 Mc/s result for Moscow is also excluded. The latter results show an exceptionally high fading time-constant with a mean of 4·7 sec and a standard deviation of 5·1; of 97 results, 6 exceeded 10 sec and the highest recorded was 41 sec. Since rapid fading is to be expected with multi-path propagation, it is understandable that for frequencies near the maximum usable frequency the single-path propagation would give slower fading.

It may be noted that, in spite of the high average σ, the spreads are considerable, and all but one of the histograms in this distance group have some values of σ below 1 sec.

(*b*) *Intermediate-Distance Asian Transmissions.*

There is a marked variation of σ with frequency and the results can be divided into two frequency groups:

15 and 18 Mc/s. Mean $\sigma = 1 \cdot 9$ sec $\pm 0 \cdot 7$ (standard deviation).
6 and 7 Mc/s. Mean $\sigma = 0 \cdot 75$ sec $\pm 0 \cdot 3$ (standard deviation).

(*c*) *Distant Transmissions.*

The fading time is most consistent, the average of 0·56 sec having a standard deviation of 0·15, with no variation over the frequency band 7–12 Mc/s.

(3.2) Diurnal Variations

The daily variations of ionospheric conditions led to the lower-frequency tests being done mainly under night conditions, so that any suggestion that the fading speed σ is dependent on time of day may also suggest a dependence upon frequency.

Fig. 8 shows the mean σ for all tests with one station at one frequency, plotted against the average local time for all the tests

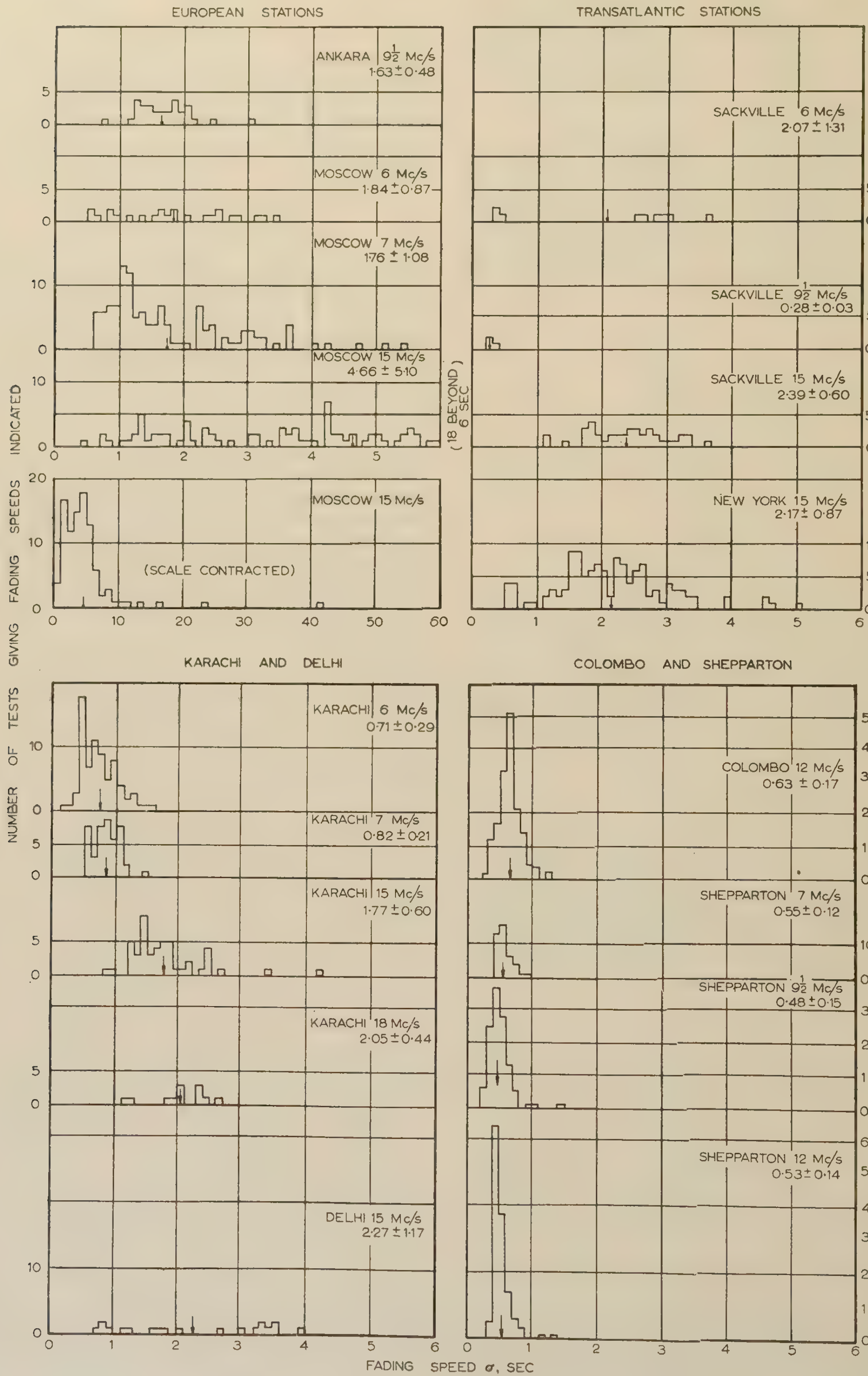

Fig. 6.—Histograms of the fading-speed parameter for various stations and frequencies.
Each unit represents one test period.

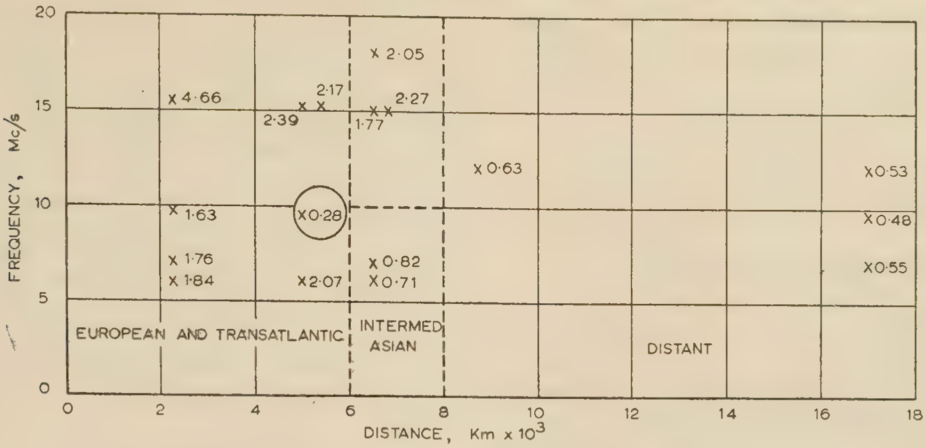

Fig. 7.—Average fading speed for all tests of a given station and frequency.

..t the mid-point of the path; the gap between the 'fast-fading' ..roup (highest mean $\sigma = 0{\cdot}82$ sec) and the slow-fading group ..lowest mean $\sigma = 1{\cdot}63$ sec) is clearly indicated. The three ..ackville $9\frac{1}{2}$ Mc/s tests (ringed) are again exceptional.

(4) GAIN DUE TO DIVERSITY RECEPTION

The advantage obtained by using a diversity system instead ..f a single-channel system depends upon four main factors:

(*a*) The quality of reception required; in telegraphy, for example, the proportion of symbols to be correctly received.
(*b*) The number n of component channels in the system.
(*c*) The fading characteristics of the individual channels.
(*d*) The degree of correlation between the fading in the individual channels.

It is possible to decide what is a reasonable lowest usable ..evel for the signal produced at the input to the receiver, depending ..pon the nature of the signal, whether it be telephony, telegraphy ..r some other form of communication, and on the various noise ..ontributions of the receiving system. The noise level will not ..ompletely define the quality of communication, for distortion ..f the waveform due to the propagation medium may also ..ontribute to the noise, even with strong signals. However, ..he strength of the signal in relation to noise is an essential ..arameter for the assessment.

The statistical approach to the diversity problem was dis..ussed by Jelonek, Fitch and Chalk,[2] who studied the equivalent ..ower gain of a diversity system. A fading signal will have an ..mplitude above the usable level for a certain proportion of the ..ime. By selecting the stronger signal at any time in a diversity ..ystem, the proportion of usable time is increased. It may be ..qually true to say that the diversity system can provide the same ..roportion of usable time as the single channel, but with less ..ower at the transmitter when diversity is used. The concept of ..ower gain is more useful to the engineer than is the time factor, ..nd Reference 2 shows how this gain may be calculated when the ..ading is independent in the two channels and when the fading ..ollows stated laws. The gain G in decibels is given by

$$G = 10 \log_{10} \left[\frac{\log (1 - \theta^{1/n})}{\log (1 - \theta)} \right] \quad . \quad . \quad . \quad . \quad (6)$$

..hen the fading obeys Rayleigh's law,

..d

$$G = \frac{20}{m}(1 - 1/n) \log_{10} 1/\theta \quad . \quad . \quad . \quad . \quad (7)$$

..en the fading obeys a power law.

..n the above equations n is the number of channels combined

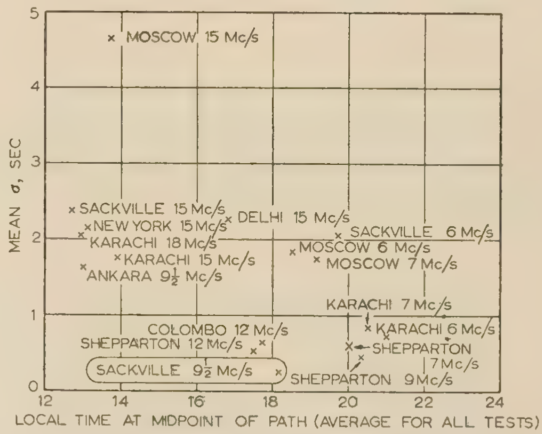

Fig. 8.—Average values of σ for a given station and frequency.

in the diversity system, θ is the proportion of unusable time that is allowed in either the diversity system or a single-channel system, and m is the index of the power law, discussed in Section 2.

The gain increases with the number of channels used, but only the two-channel case is dealt with here. A triple system is often used with spaced-aerial diversity, but the polarization system normally utilizes only the vertical and horizontal components of the wave. Evaluation experiments were also simpler to arrange with only two channels.

(4.1) Correlation Between Fading in a Diversity System

The instantaneous amplitudes of the two signals from which a diversity output is obtained will not, in general, be unrelated; the interdependence is measured by the correlation coefficient R, where $+1 \geqslant R \geqslant -1$. When $R = +1$, the two amplitudes vary together; when $R = 0$, the signals are said to be independent. If $R = -1$, an increase in one amplitude corresponds to a decrease in the other. For maximum gain in the diversity system, R should be negative and ideally be equal to unity; unfortunately, negative correlation is rarely found in practice.

For independent signals ($R = 0$), the proportion of time that the two signals spend simultaneously below level S is the product of the corresponding proportion for each individual signal, whatever the fading law. This is shown for Rayleigh fading by the curve marked $R = 0$ in Fig. 3, and is, in fact, the function $\{1 - \exp [-0{\cdot}693(S/S_m)^2]\}^2$ plotted against S.

As shown in Fig. 3, for correlation greater than zero the

proportion of time simultaneously below level S is greater, until, at $R = 1$, the curve coincides with the 'single channel' curve, with no diversity gain.

The statistical law giving these curves is

$$P(S) = \int_0^S \int_0^S \frac{xy}{1-\rho^2} \exp\left[-\frac{x^2+y^2}{2(1-\rho^2)}\right] I_0\left(\frac{\rho xy}{1-\rho^2}\right) dxdy$$

where ρ is related to the correlation coefficient R by

$$R \simeq \frac{\pi\rho^2}{4(4-\pi)}$$

for values of ρ up to 0·5.

In fact, ρ is the correlation of the Gaussian distributions mentioned above. I_0 is the Bessel function of the first kind of imaginary argument and order zero, and x and y are normalized instantaneous values of the signal levels, S' and S'', related to the median values S_m by

$$x = 1{\cdot}18\frac{S'}{S_m} \qquad y = 1{\cdot}18\frac{S''}{S_m}$$

The Rayleigh fading law has been assumed, and the curves shown in Fig. 3 are known as bivariate probability integrals. The assumption of the Rayleigh law implies that the correlation between the signals in the individual channels cannot be negative; this theoretical deduction agrees with what has been found experimentally.

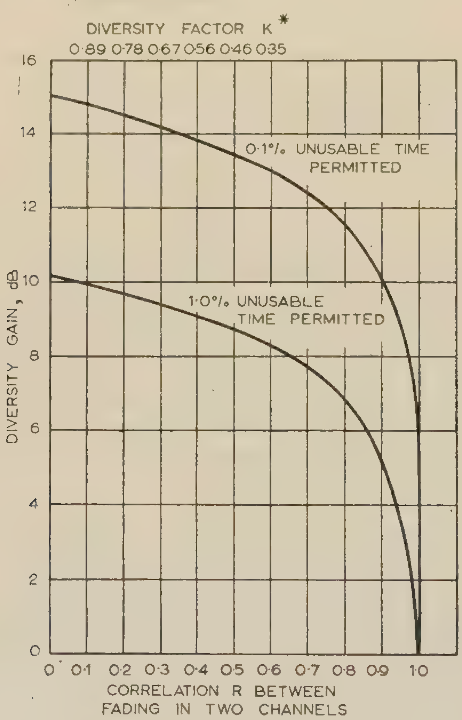

Fig. 9.—Double diversity gain as a function of correlation between signals, assuming Rayleigh law fading.

* Calculated from the relation $K \simeq 1 - 1{\cdot}09R$ for R not large.

The effect of correlation on diversity gain is shown in Fig. 9, where the gain G is plotted as a function of R for two values of θ, the proportion of unusable time permitted in the particular circumstances. In practice, θ is a small fraction and 0·1% and 1% are typical of operating requirements. The gain is evaluated for Rayleigh-law fading. Thus, Fig. 3, a line through $\theta = 0{\cdot}01$ parallel to the abscissa meets the single-channel curve at $S = -18{\cdot}4$ dB, and the $R = 0$ diversity curve at $S = -8{\cdot}2$ dB, giving a diversity gain for $R = 0$ of 10·2 dB.

It can be seen from Fig. 9 that G decreases slowly with increasing R, and is within 2 dB of the maximum value when R is as high as 0·6. The gain is far more dependent on the unusable time permitted (θ); the more reliable the communication circuit is required to be (i.e. θ smaller), the greater is the diversity power gain. This holds good for fading laws other[2] than the Rayleigh law. An assumption of independent fading in calculating the diversity gain is justified for correlation less than about 0·6.

For the Rayleigh law, a simple relation exists between the proportions A and B of the total time that the individual signals fall below a fixed level, the proportion C that both simultaneously fall below this level, and the correlation coefficient R. It is given by

$$\frac{AB}{C} = K \simeq 1 - 1{\cdot}09R \quad . \quad . \quad . \quad . \quad . \quad (8)$$

provided that A and B are fairly small fractions and that the correlation R is less than about 0·5.

The diversity factor, K, is itself a figure of merit of the diversity system, and the apparatus to be described measures A, B and C directly, so that K may be easily calculated. For $K = 1 (R = 0)$ the signal strengths are independent of one another; if $K > 1$ a negative correlation is implied; and for $K < 1$, corresponding to most of the experimental measurements, the signal levels are positively correlated and the diversity gain is correspondingly reduced.

The statistical method used to compute the correlation coefficients for the diversity systems used in the measurements assumed that the amplitude distribution followed the Rayleigh law. The experiments gave the proportions of time faded, A and B, by the two signals below level S_1 and S_2, respectively, and also the proportion of time C that both faded simultaneously. The normalized levels S_1/S_{m1} and S_2/S_{m2} were then found by using the Rayleigh-law relations

$$A = 1 - \exp\left[-0{\cdot}693(S_1/S_{m1})^2\right] \quad . \quad . \quad . \quad (9)$$

$$B = 1 - \exp\left[-0{\cdot}693(S_1/S_{m2})^2\right] \quad . \quad . \quad (10)$$

Tables were drawn up giving the correlation coefficient R with the normalized levels S_1/S_{m1} and S_2/S_{m2} and the fraction C as arguments. Hence, from the measured fractions A, B and C, a value of the correlation was deduced.

In the particular case when the signals in the two channels have the same median values, the curves of Fig. 3 show the proportion of time faded below the datum level, the datum level and, for the diversity case, the correlation. For example, if the simultaneous fraction $C = 0{\cdot}04$ and $S_1 = S_2 = 6$ dB median value, this corresponds to a correlation of about 0·4 between the two signals.

The results of all the tests are shown as histograms in Fig. 10, the abscissa giving the correlation range and the ordinate showing the number of times that this correlation occurred. The results are discussed in Section 7.

The Rayleigh law permits only zero or positive correlation between the two individual signals, whereas the experimental results sometimes gave negative correlation. These results were thought to be caused by sampling fluctuations.

Sampling theory was used to estimate the significance of the correlations deduced from the measurements, and it was possible to do this even for apparently negative correlations to which no numerical value could be put. The tests were applied to many of the measurements, including the whole of the first set made at Baddow, and, with two exceptions, the negative corre-

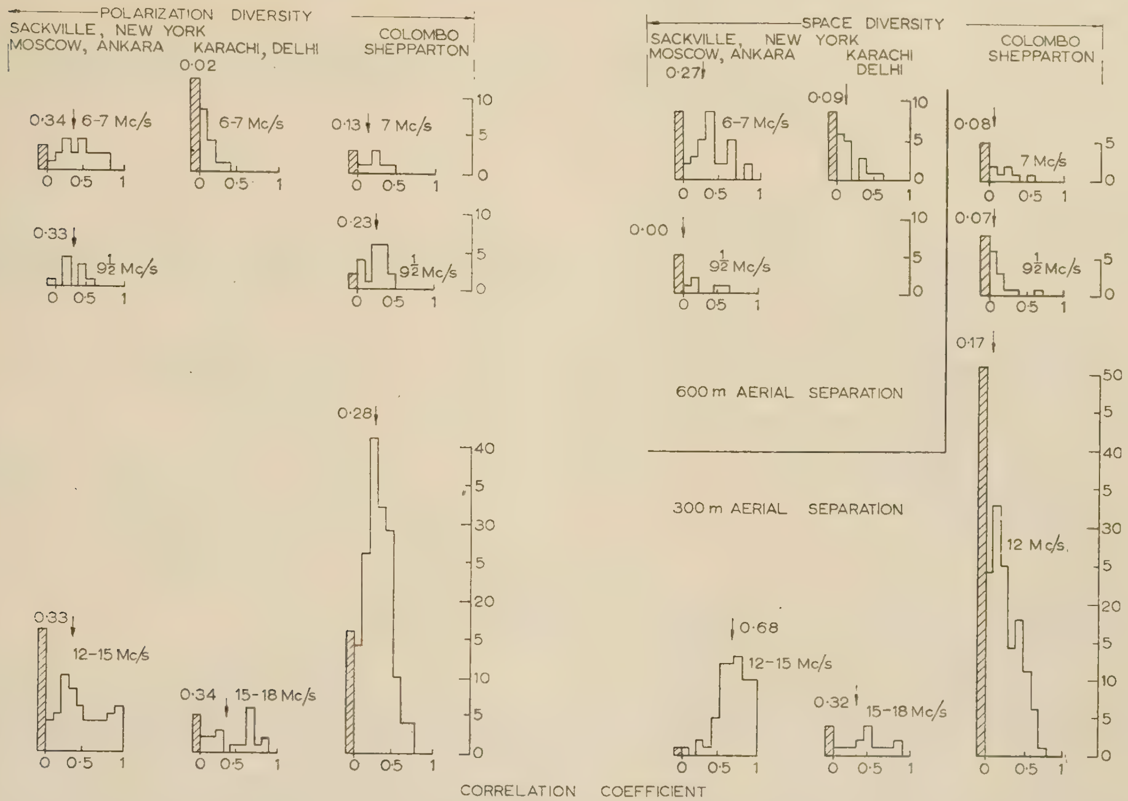

Fig. 10.—Correlation histograms of results from different groups of stations, with polarization and space diversity.
Each result represents one test period. Hatching denotes negative correlation.

lations were found not to be significant. Since one derivation of the significance test did not assume a Rayleigh distribution of amplitude, the fact that almost no negative correlations were significant provides indirect support for the Rayleigh hypothesis. An appreciable number of significant negative correlations would mean that this hypothesis would have to be abandoned.

Since no numerical value could be put to the negative correlations, the mean of the results was estimated by treating as zero the negative correlations and an equal number of the smallest positive correlations, thus avoiding the introduction of bias into the mean.

(4.2) The Space Structure of the Fading

In Section 3 it is shown that a fading signal may be given a time-characteristic σ derived from the autocorrelation of the fading curve. In the same way a characteristic spatial structure size may be obtained for the wave received.

The space correlogram is defined by an equation similar to eqn. (4) with distance d taking the place of τ as variable. Similar remarks about the form of the functional dependence on d apply, and in order to make use of the limited data it is necessary to assume a Gaussian relation which may be written

$$\rho(d) = \varepsilon^{-d^2/2X^2} \quad . \quad . \quad . \quad . \quad . \quad (11)$$

where X is the distance between points at which the correlation of the fading falls to 0·61.

The correlation $\rho(d)$ is that existing at a particular instant between the signal amplitude at two points separated by a distance d, and it is, of course, the correlation inferred from the space-diversity measurements. Knowing the separation of the aerials, and taking the mean value of correlation measured in tests with these aerials, a mean structure size X may be obtained. This structure size is a property of the incoming wave.

It is possible to compare polarization diversity and dual space diversity by estimating from the measurements the separation of the spaced aerials which would give the same mean correlation as that measured for polarization diversity. For example, if the polarization diversity gave a correlation of 0·61, this would correspond to space diversity with the aerials separated by the structure size X. The experimental measurements have been compared in this way and results are given in Section 7.

(5) EXPERIMENTAL METHODS

(5.1) Comparison of Diversity Systems

The object of the measurements was to compare polarization- and space-diversity systems and to evaluate the gains of the diversity systems over single-channel working. A direct comparison was made by integrating the time intervals that each individual signal spent below a set datum level and comparing this with the time spent below the same level by both signals simultaneously. The number of excursions of the signal below the datum level was also counted to give the average length of a fade.

The apparatus used could handle four individual signals, and the measurements were made with two polarization-diversity aerials and two spaced aerials simultaneously. The polarization signals were derived from aerials erected in the same position but responsive to differently-polarized components of the incident wave; the space signals were derived from two aerials similarly orientated but separated in space by at least 270 m.

Another series of measurements was made to determine the

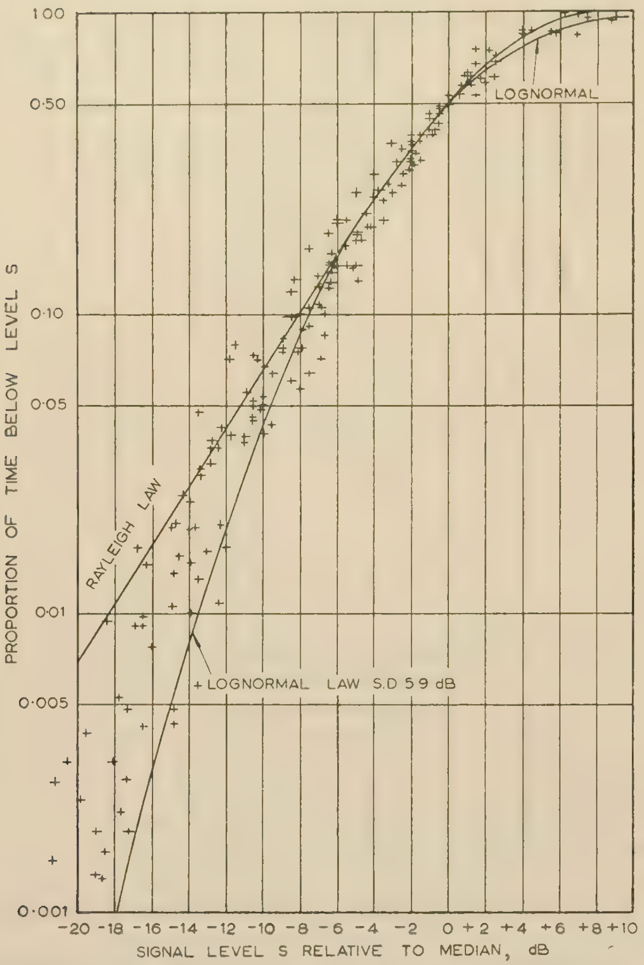

Fig. 11.—Experimental measurement of the fading characteristics of single-frequency signals.

Full line indicates Rayleigh distribution.

length of time that a single signal spent below each of four preset levels, i.e. to find four points on the probability-integral curve. The points on Fig. 11 are the results of these measurements.

(5.2) Receiving Aerials

For polarization diversity, dipole aerials were arranged in a vertical plane and at right angles to each other. They faced approximately east–west, since the majority of broadcasting stations lie in these directions. The space-diversity aerials were half-wave dipoles, or, in some of the higher-frequency measurements made at Brentwood, horizontal arrays of dipoles (h.a.d.). Separations ranged from 309 to 700 m. Wide-band transformers were used to match the impedances of aerials and coaxial cables feeding the signals to the receivers.

Where practicable, a check of the relative sensitivities of each of a pair of aerials was made, using a local oscillator some wavelengths from the aerials. The pick-up was compared for different polarizations of the incident wave. For the frequency range 12–17·5 Mc/s, the aerials were 10 m long; for 6–9 Mc/s, the aerials were 25 m. The h.a.d. aerials were chosen to suit the frequency and direction of the received transmission. The arrangement of dipole aerials is shown in Figs. 12 and 13.

Over the range of frequencies used, the dipole aerial sensitivities, including feeder losses, varied less than 5 dB, and for most measurements the difference was less than this. For the

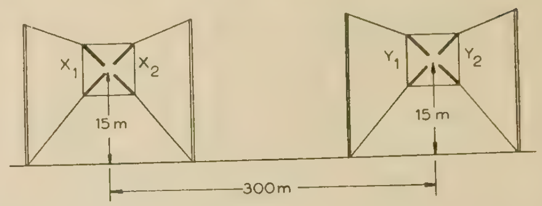

Fig. 12.—Dipole aerials erected on sites about 300 m apart.

X_1 and X_2, Y_1 and Y_2 are mutually at right angles and inclined at 45° to the horizontal. (Chelmsford aerials.)

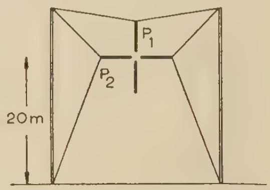

Fig. 13.—Dipole aerials, P_1 vertical and P_2 horizontal.

h.a.d. aerials the greatest difference was 8–10 dB, but for most pairs the difference was lower.

(5.3) Measuring Technique

Fig. 14 shows the block diagram of the apparatus and the circuit from the output of the receivers to the counters. Short pulses of known repetition rate were passed through a gating valve which was opened when the signal fell below the predetermined level. The gated pulses were applied to mechanical counters, which recorded the number passed during a test period; by dividing the counter reading by the total number of pulses during the test period, the fraction of the time during which the signal was below the datum level could be calculated. Coincidence circuits also measured the time during which the two signals simultaneously had amplitudes less than the datum levels of the individual channels.

Each time that one of the gating valves opened, a further mechanical counter advanced one digit, thus counting the number of fades in each channel and enabling the average duration of a fade to be computed.

From Fig. 14 it can be seen that the signals were passed at intermediate frequency from the receivers to the measuring apparatus, where they were amplified and rectified. Following the detectors were low-pass filters with 20 c/s cut-off, which removed audio modulation, giving rectified outputs proportional to the amplitudes of the received carriers. Trigger valves operated the gates which allowed the pulses to pass to the counters.

In making probability-integral recordings, the four channels of the recording apparatus were all connected to a single receiver; the gain of the channels was attenuated in steps to give equal increments between the four counting levels. The actual level at which each counter operated was determined by signal-generator measurement.

Consideration was given to the duration of a recording. A graph of signal amplitude as a function of time shows, besides rapid deep fading, slow changes occurring over periods of many minutes. Only the effects of rapid changes of level are reduced by diversity reception, so that the present measurements should exclude the slower variations. The latter can in effect be smoothed out by sampling the signal over a period short enough for the median value to be considered constant during the recording, the datum level being adjusted between recording periods if necessary. This sets an approximate upper limit to the duration

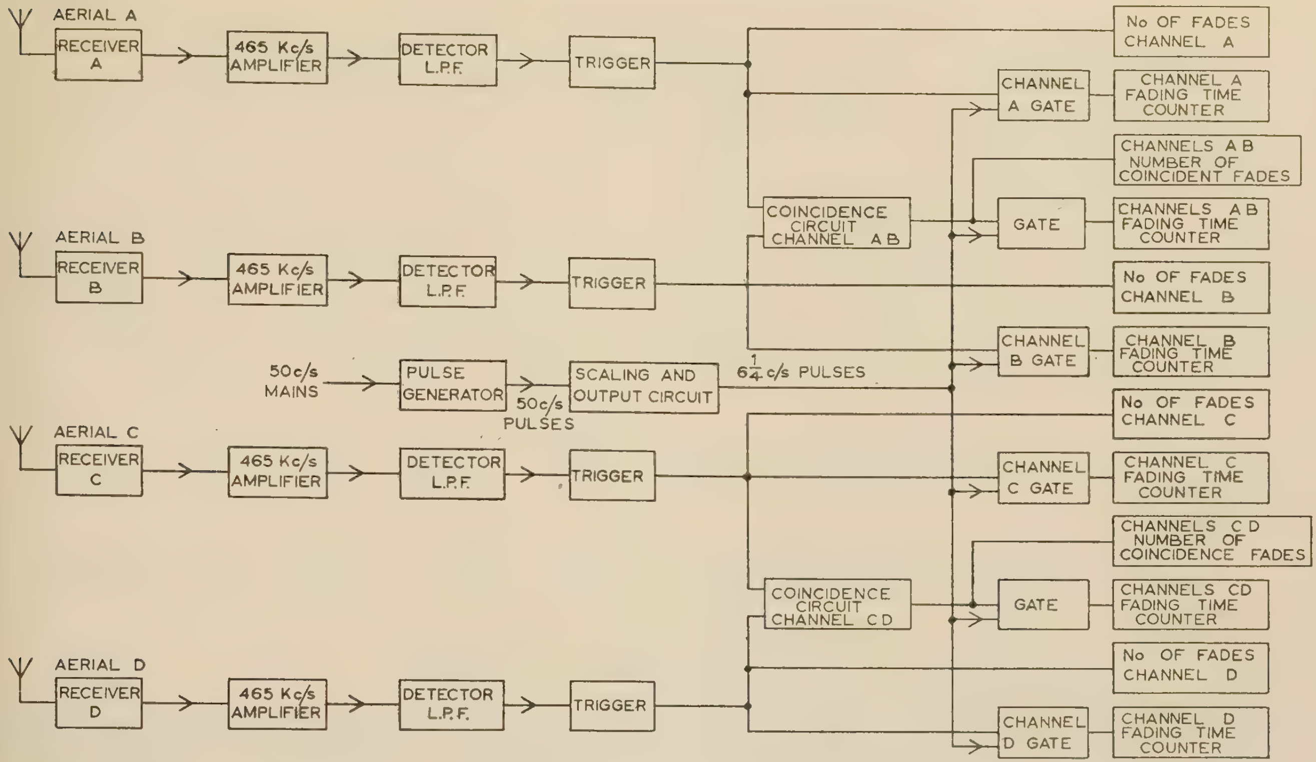

Fig. 14.—Arrangement of circuits for diversity-aerial comparison.

of a test period, which can be estimated from pen recordings over long periods. This procedure enabled the simplifying assumption of constant median value for any one test period to be made in subsequent analysis of the experimental data.

The lower limit for the test period was set by the speed of the fast-fading component and by the method adopted for measuring time by integrating bursts of pulses having a known repetition rate. A statistical study of the problem showed that the estimate of the time interval had a proportional standard deviation of $(2/p)^{1/2}$, where p is the number of pulses counted, provided that the average fade contained more than two pulses. Thus, for 100 pulses counted, the time would have a standard deviation of 14%. In the experiments the measurement period was between 6 and 12 min, with the majority of 8 min.

The level at which the signal amplitude was sampled was also subject to the above considerations. On the average, the counting level was set so that the amplitude was less than the datum for about 20% of the time; the time for coincident fading was then usually greater than 5%. This was regarded as a satisfactory compromise although it was often found difficult to achieve in practice owing to the ever-present slow component of the level variation.

(6) EXPERIMENTAL RESULTS

The first group of measurements was made near Chelmsford and the second and third groups were obtained with aerials at Brentwood Wireless Receiving Station. Broadcast transmissions were used, mostly on frequencies of about 6, 7, 9, 12 and 15 Mc/s. Transmission paths ranged from 2250 km (Moscow) to 17 000 km (Shepparton, Australia).

After each recording period of about 8 min, the proportion of time that the signal was less than the datum level was calculated from the counter readings, for each channel individually and for the simultaneous fading. Each channel was adjusted for a test as far as practicable to the same datum level in terms of field strength; this was done by using a standard signal generator to set the receiver gains, taking into account the differences in aerials and feeders, usually less than 5 dB in all.

As a practical criterion by which to compare simply the simultaneous performance of double space and polarization diversity, the factor $K = AB/C$ was evaluated for each system (Section 5). The higher the value of K, the better the diversity effect ($K = 1$ for independent fading in the two channels). The scatter in K is indicated by the standard deviation for all the test periods, and the number of measurements is also entered to give an idea of the relative weight to be attached to each result. Since all signals were sampled at approximately the same level, the merit of one diversity system over the other was shown to be a difference in the diversity factor K.

(6.1) Measurements at Chelmsford

The first set of measurements was made at Chelmsford on frequencies of 12 and 15 Mc/s from May to September, 1953. Two pairs of aerials, half-wave dipoles at a frequency of 15 Mc/s, were erected 300 m apart (Fig. 12). Both aerials of each pair lay in the same vertical plane, were mutually at right angles and inclined at 45° to the horizontal. Their common centre was about 15 m above the ground and their common plane lay in a north–south direction, so that transmissions from east and west could be used. Each pair of aerials X_1X_2 and Y_1Y_2 could be used to give polarization diversity, and similarly orientated pairs X_1Y_1 and X_2Y_2 could be used for spaced-aerial diversity with 300 m separation. In practice, duplicate sets of measurements were made from the four dipoles, since enough coincidence-counter circuits were available to record the simultaneous fading in X_1X_2, Y_1Y_2, X_1Y_1 and X_2Y_2.

The measurements are summarized in Table 1. In general the diversity factor K is less than unity, indicating that the fading in the pairs of aerials is positively correlated.

Table 1

FIRST SET OF MEASUREMENTS AT CHELMSFORD, 11TH MAY TO 25TH SEPTEMBER, 1953

Station	Distance, km	Frequency, kc/s	Diversity factor K, standard deviation and (in parentheses) number of test periods averaged						Diversity gain,* G, dB, and (in parentheses) correlation coefficient	
			Polarization diversity			Space diversity				
			Aerials X_1 and X_2	Aerials Y_1 and Y_2	Average	Aerials X_1 and Y_1	Aerials X_2 and Y_2	Average	Polarization diversity	Space diversity
Shepparton, Australia	17 000	11 900	0·77 ± 0·18 (9)	0·77 ± 0·08 (9)	0·77 (18)	1·00 ± 0·17 (9)	0·96 ± 0·19 (9)	0·98 (18)	14·4 (0·25)	14·9 (0·09)
Colombo, Ceylon	8 700	11 975	0·69 ± 0·22 (82)	0·79 ± 0·26 (82)	0·74 (164)	0·90 ± 0·36 (83)	0·89 ± 0·29 (83)	0·90 (166)	14·3 (0·29)	14·6 (0·19)
Delhi, India	6 800	15 085 15 380	0·83 ± 0·45 (12)	0·54 ± 0·38 (12)	0·73 (32)	0·58 ± 0·30 (12)	0·68 ± 0·27 (13)	0·72 (33)	13·9 (0·39)	13·9 (0·40)
Karachi, Pakistan..	6 500	15 335	0·91 ± 0·13 (4)	0·80 ± 0·14 (4)		1·01 ± 0·30 (4)	1·09 ± 0·40 (4)			
Schenectady, WGEO, WRUL, U.S.A.	5 500 5 300	15 330 15 200	0·88 ± 0·15 (17)	0·95 ± 0·37 (17)	0·91 (34)	0·60 ± 0·16 (17)	0·67 ± 0·21 (17)	0·63 (34)	14·7 (0·17)	13·1 (0·60)
Moscow, U.S.S.R.	2 250	11 765 11 970	0·54 ± 0·10 (5)	0·43 ± 0·11 (5)	0·56 (26)	0·63 ± 0·24 (5)	0·65 ± 0·17 (5)	0·59 (26)	12·6 (0·66)	12·9 (0·62)
		15 100	0·41 ± 0·19 (8)	0·79 ± 0·19 (8)		0·36 ± 0·22 (8)	0·76 ± 0·27 (8)			

* Diversity gain is calculated assuming Rayleigh fading law and a permissible time loss of 0·1%.

The diversity gain as defined in Sections 4 and 5 is calculated in the last two columns; for this purpose the fading is assumed to follow a Rayleigh law, and a permissible time loss of 0·1% is also assumed. The calculation takes into account the correlation between the fading of the individual signals; the correlation coefficient (in parentheses in the last two columns) is estimated from the 'fraction faded' measurements by referring to tables of the probability integral, as explained in Section 5.

If the fading were independent, the diversity gain for 0·1% time lost would be 15·1 dB, and it can be seen that the experimental figures are within 2·5 dB of this figure, even for the most highly correlated fading from Moscow ($K = 0·56$). This deficiency is small compared with the actual gain over non-diversity reception, but the assumption of independent fading in a gain prophecy would give an optimistic answer.

The transmitting stations are arranged in order of decreasing distance, and it can be seen that for space diversity there is a tendency for the diversity factor to increase with distance, and for the longest distances (Colombo and Shepparton) the fading is almost independent. The variation with distance for polarization diversity is not evident; for the two longest paths the diversity factors are less than those for space diversity, but a marked difference occurs for the American transmissions, leading to an estimated 1·6 dB in favour of the polarization system. The measurements for Moscow transmissions show that the fading was far from independent for both systems and on average there was little difference in the diversity gain.

These measurements may be summarized by saying that polarization diversity gave a gain within 0·5 dB of the gain from spaced aerials at 300 m separation, and, for transmissions from Schenectady, polarization diversity had an advantage of about 1·6 dB.

(6.2) Measurements at Brentwood (12–18 Mc/s)

The second set of measurements was made at Brentwood Wireless Receiving Station during April and May, 1954, chiefly on frequencies of about 12 and 15 Mc/s, with a few at 17·5 Mc/s. The polarization-diversity aerials consisted of one pair of the dipoles used for the previous measurements, which were, however, rigged so that one was vertical (P_1) and one horizontal (P_2) with the common centre 20 m above the ground (Fig. 13). The vertical plane was in the north–south direction as before.

The aerials therefore responded to the vertical and horizontal polarization of the incident radiation. Two spaced horizontal arrays of dipoles were used for spaced-aerial diversity, the particular pair being chosen to suit the frequency and direction of transmission; the separation was 300 m in all cases. The sensitivities of the receivers were reduced to allow for the gain of the arrays over the dipoles. Table 2 gives the results in the same form as Table 1. As before, the fading in the two channels is not independent, but in this case the polarization system gives slightly better diversity than the spaced aerials; the difference between the systems is once again greatest for the transatlantic paths. For the shortest path (Moscow) there were great differences between the values of diversity factor. The gain for polarization diversity was very close to that for independent signals (15 dB); for spaced-aerial diversity the gain was always less than with the polarization system, the difference reaching 3 dB for the Canadian transmissions.

Comparing Tables 1 and 2 for the same stations it can be seen that the corresponding values of the space diversity factor are less in the second group, although the aerial spacings were in each case about 300 m. The difference might be related to site or season, or to aerial directivity. With polarization diversity no clear variation is apparent.

The percentage of time faded, recorded from the vertical aerial P_1 and the horizontal aerial P_2, have been summarized in Table 3. The maximum difference was for Shepparton (12 Mc/s) transmissions; assuming the Rayleigh fading law, the time difference indicated that in this case the average signal was 2·5 dB greater in the horizontal aerial than in the vertical.

(6.3) Measurements at Brentwood (6–10 Mc/s)

In the third set of measurements taken at Brentwood, the frequency range 6–9·5 Mc/s was used. The polarization-diversity aerials were similar to those shown in Fig. 12, cut to half-wavelength at 7 Mc/s and centred 20 m above the ground. Space-diversity aerials were horizontal half-wave dipoles up to

Table 2

SECOND SET OF MEASUREMENTS AT BRENTWOOD WIRELESS RECEIVING STATION, 7TH APRIL TO 7TH MAY, 1954

Station	Distance, km	Frequency, kc/s	Mean diversity factor K, standard deviation and (in parentheses) number of test periods averaged		Average diversity gain G, dB and (in parentheses) correlation coefficient	
			Polarization diversity	Space diversity	Polarization diversity	Space diversity
Shepparton, Australia	17 000	11 900	0·83 ± 0·13 (17)	0·78 ± 0·23 (18)	14·5 (0·21)	14·5 (0·21)
Karachi, Pakistan	6 500	17 770 15 335	0·84 ± 0·07 (5)	0·60 ± 0·25 (6)	14·5 (0·21)	13·9 (0·38)
WRCA, U.S.A.	5 500	15 130	0·84 ± 0·16 (25)	0·50 ± 0·19 (25)	14·6 (0·20)	13·0 (0·60)
Sackville, Canada	4 500	15 090 15 190	0·96 ± 0·30 (10)	0·38 ± 0·18 (10)	15·1 (0)	12·2 (0·72)
Moscow, U.S.S.R.	2 250	15 180 15 220 15 270	0·69 ± 0·50 (22)	0·48 ± 0·22 (23)	13·5 (0·49)	* (0·95)

* Not computed because many results were outside the range of the Rayleigh probability-integral tables.

Table 3

COMPARISON OF AVERAGE SENSITIVITY OF HORIZONTAL AND VERTICAL AERIALS

Station	Frequency, kc/s	Total recording time, min	Number of tests	Percentage time below reference level		Calculated ratio of horizontal/vertical signal at input, dB
				Vertical aerial P_1	Horizontal aerial P_2	
Shepparton, Australia ..	11 900	120	14	22·0	12·9	2·5
WRCA, U.S.A. ..	15 130	191·5	25	29·1	23·7	0·9
Sackville, Canada ..	15 090 15 190	94·25	12	15·2	15·3	0
Moscow, U.S.S.R. ..	15 180 15 220 15 270	180	22	15·3	18·2	−0·8

Table 4

THIRD SET OF MEASUREMENTS AT BRENTWOOD WIRELESS RECEIVING STATION, 7TH OCTOBER TO 4TH DECEMBER, 1953

Station	Distance, km	Frequency, kc/s	Mean value of diversity factor K, standard deviation and (in parentheses) number of test periods averaged		Average diversity gain G, dB, and (in parentheses) correlation coefficient	
			Polarization diversity	Space diversity	Polarization diversity	Space diversity
Shepparton, Australia ..	17 000	7 220 9 580	0·88 ± 0·13 (13) 0·76 ± 0·11 (35)	0·89 ± 0·19 (13) 0·97 ± 0·18 (35)	14·5 (0·24)	14·9 (0·09)
Karachi, Pakistan	6 500	6235 7 010	1·03 ± 0·18 (21) 0·99 ± 0·22 (15)	0·97 ± 0·26 (22) 0·93 ± 0·22 (15)	14·9 (0·06)	14·9 (0·10)
Sackville, Canada	4 500	6 060 9 610	0·82 ± 0·18 (3) 0·86 ± 0·15 (2)	0·61 ± 0·35 (4) 0·76 ± 0·37 (2)	14·2 (0·32)	13·8 (0·43)
Ankara, Turkey	2 800	9 515	0·75 ± 0·14 (10)	1·05 ± 0·39 (10)	14·2 (0·32)	14·9 (0·08)
Moscow, U.S.S.R.	2 250	6 195 7 200 about	0·86 ± 0·35 (8) 0·79 ± 0·28 (37)	0·55 ± 0·21 (8) 0·85 ± 0·28 (34)	13·9 (0·39)	14·4 (0·25)

7 Mc/s and h.a.d. aerials for the 9 Mc/s band. The spacing ranged from 400–1 000 m with most measurements at about 700 m separation.

Results are presented in Table 4 in the same form as before. The previous connection between spaced-aerial diversity-factor and distance is not evident. All diversity gains are within 1·3 dB of the 15 dB gain for independent fading.

(7) ASSESSMENT OF RESULTS

Excluding the measurements relating to Moscow, there is little variation of the diversity factor for the polarization system with frequency, length of path or direction, the range of mean values being 0·73–1. The Moscow transmissions gave a mean diversity factor of 0·8 at 7 Mc/s, falling to 0·6 at 15 Mc/s for the first set of measurements.

Spaced aerials gave diversity factors in the range 0·7–1 for all transmissions except those from Moscow and North-East America, where the factor is lower at all frequencies; transmissions from Canada on about 15 Mc/s gave the lowest K value of 0·38. That for Moscow falls from 0·8 at 6–7 Mc/s to 0·5 for one set of observations at 15 Mc/s, but in these cases the aerial separation also changed from 600 m to 300 m.

Means of the results of all the experiments are summarized

Table 5

MEAN VALUES OF ALL COMPARISONS OF SPACE AND POLARIZATION DIVERSITY

Station	Distance, km	Approximate frequencies, Mc/s	Mean diversity factors		Spaced aerial separation, m
			Polarization diversity	Space diversity	
Shepparton, Australia	17 000	7	0·78 (83)	0·92 (84)	400
		9·5			300
		12			270
Colombo, Ceylon	8 700	12	0·74 (164)	0·82 (168)	270
Delhi, India, ...	6 800	6	0·88 (73)	0·82 (76)	
and		7			600
Karachi, Pakistan	6 500	15			300
		17·5			270
U.S.A.	5 300	6	0·89 (74)	0·56 (75)	600
and		9·5			300
Sackville, Canada	4 500	15			270
Moscow, U.S.S.R.	2 250	6	0·71 (93)	0·65 (91)	600
		7			300
		12			270
		15			

in Table 5; the figures in parentheses are the number of results included in a mean.

From the standpoint of gain calculation, the diversity signals could usually be assumed to be independent; the exception is the spaced-aerial diversity system, with transmissions from North America or Moscow (or a similar distance), when a correlation parameter should be used in a gain calculation.

All the individual experimental results have been examined from a statistical point of view. The correlation coefficients of the fading of the signals on the diversity-aerial system have been calculated for the test results in the way described in Section 5. Fig. 10 shows histograms of the correlations; the ordinate gives the number of test periods during which the correlation was within a 0·1 interval in the correlation range, which is plotted as abscissa. The histograms are divided into three main groups, namely polarization diversity, space diversity with 300 m spacing and space diversity with 600 m spacing.

Each group is divided vertically according to frequency, and horizontally according to distance; the characteristics of the fading indicate that the transmissions used for the tests fall into three broad categories of distance.

(*a*) *European and Transatlantic Transmissions: Moscow, Ankara, New York, East Canada.*

Space diversity (300 m). Mean correlation = 0·68 (12–15 Mc/s).
Space diversity (600 m). Mean correlation = 0·15 (6–9½ Mc/s).
Polarization diversity. Mean correlation = 0·33 (all frequencies).

From the space-diversity correlations with different aerial separations the mean structure size X was obtained (see Section 5). This structure size was 330 m for the results with 300 m separation, and 300 m with 600 m separation.

On the basis of this structure size, the correlation of the fading when using polarization diversity would correspond to a space-diversity system with 480 m between the aerials.

(*b*) *Asian Transmissions: Intermediate distances, Karachi, Delhi.*

Space diversity (300 m). Mean correlation = 0·32 (15–18 Mc/s).
Space diversity (600 m). Mean correlation = 0·09 (6–7 Mc/s).
Polarization diversity. Mean correlation = 0·14 (all frequencies).
Structure size = 200 m for 300 m aerial separation.
Structure size = 270 m for 600 m aerial separation.

The significance of the results with 600 m spacing was small, particularly after adjusting for negative values; the weighted mean structure size was put at 210 m. Polarization-diversity results now correspond to 420 m spacing for space diversity.

(*c*) *Longest Distances: Colombo and Australia.*

Space diversity (300 m). Mean correlation = 0·15 (7–12 Mc/s).
Polarization diversity. Mean correlation = 0·27.
Structure size = 150 m for 300 m aerial separation.
Polarization diversity is equivalent to an aerial separation of 240 m.

As the distance increases, the spatial structure size becomes less and the aerial separation equivalent to polarization diversity is also less. Except for the most distant stations, polarization diversity gave better results than space diversity with 300 m separation, but worse than that with 600 m separation. For the greatest distances the correlation is so small that there will be little practical difference between space and polarization diversity (see Fig. 5).

For space diversity there is no evidence of a connection between fading correlation and frequency within the range 6–18 Mc/s. It can be seen from Fig. 10 that for polarization diversity (with the more distant stations) there is a trend towards higher correlation at the higher frequencies; the range of correlation is 0·02 at 7 Mc/s to 0·34 at 15–18 Mc/s for intermediate distances, and 0·13 at 7 Mc/s to 0·28 at 12 Mc/s for the longest distances.

In discussing the relative merits of the systems in terms of correlation coefficient, the relation between correlation and gain shown in Fig. 9 should be remembered; even a change from 0·6 to zero in correlation would be equivalent to only 2 dB of power gain.

(8) EFFECT OF DIVERSITY ON TELEGRAPH DISTORTION

As it was suggested that polarization diversity could produce distortion of telegraph signals, a practical comparison of two systems was made by measuring the incidence of telegraph distortion with a double-diversity receiver, using the spaced and polarization aerials alternately for periods of a few minutes. The measuring apparatus did not take into account the effect of amplitude fading, the telegraph distortion being measured only when the signal exceeded a minimum value.

Transmissions from Barbados on frequencies 5 305 and 15 930 kc/s were received. The transmitter was keyed at 50 c/s, either on-off or frequency shift, and the telegraph output signal was fed into the distortion measuring apparatus, which measured the number of characters having more than a pre-set amount of distortion; this setting was continuously adjustable between 0 and 25%. Test periods ranged from 2 to 7 min, a period on one aerial system being immediately followed by an equal

Table 6

Effect of Diversity System on Distortion of Telegraph Signals, Excluding the Effect of Drop-Out of Signal Due to Amplitude Fading

Date	Frequency, kc/s	Transmission and keying	Diversity aerials used		Distortion setting, %	Total duration of tests, min	Number of tests	Incidence of distorted characters (%)	
			Polarization	Space				Polarization	Space
4th–5th February, 1954	5 305	50 c/s 'reversals' on–off keyed	Dipoles mutually perpendicular (See Fig. 12)	Horizontal dipoles B_1 and B_2 600 m apart	20	46	17	0·8	1·6
					15	57	19	2·8	3·6
					12·5	2	1	2·6	4·8
						105	37	2·1	3·0
				Horizontal arrays of dipoles 5 and 6 400 m apart	20	3	1	0·28	0·35
					15	2	1	2·75	3·2
						5	2	1·3	1·5
15th–16th February, 1954	5 305	50 c/s 'reversals' on–off keyed	Dipoles mutually perpendicular (See Fig. 12)	Horizontal dipoles B_1 and B_2 600 m apart	25	28	12	1·05	0·9
					20	28	12	1·2	1·05
					15	20	10	1·2	1·35
					10	16	8	4·95	4·90
						92	42	1·85	1·74
25th March, 1954	15 930	50 c/s 'reversals'. Frequency shift keyed	Dipoles mutually perpendicular. One vertical, one horizontal (See Fig. 13)	Horizontal arrays of dipoles 15 and 16 500 m apart	7·5	34	6	0·027	0·039
					5	35	10	0·23	0·18
					2·5	13	7	2·57	2·54
						82	23	0·52	0·50

period on the other. The diversity receiver automatically selected the signal giving the greater amplitude.

Average results over several hours' measurement showed little difference in the incidence of distorted characters between the two systems. Results are summarized in Table 6.

(9) CONCLUSIONS

(*a*) Polarization diversity gave a gain about equal to that from a double space-diversity system with dipole aerials separated by at least 300 m. The polarization system is to be preferred for transmissions from Canada and the United States.

(*b*) The fading of individual signals used in a diversity system is not usually independent, and when the correlation is high its value should be taken into account in estimating the gain due to diversity working.

(*c*) For a given fractional time-loss, the values calculated suggest upper limits for the diversity gain, but the improvement achieved in practice is bound up with the design of the receiving equipment.

(*d*) The better the quality required of a system in terms of fractional time below a stated level of signal, the greater the advantage of using diversity reception. When diversity reception is used there is little to be gained by using either extremely large aerial spacings or space diversity in place of polarization diversity.

(*e*) The experimental evidence confirms that the rapid fading of h.f. signals is close to the Rayleigh law.

(*f*) Throughout the wide range of frequencies and distances used, the spatial characteristic, X, of the fading always falls in the range 150–330 m, and it varies more with distance than with frequency.

(*g*) Aerials separated by the spatial characteristic distance should provide diversity gain within 2 dB of the optimum.

(*h*) The practical implementation of polarization diversity for high-quality circuits must await the design of satisfactory directional aerials for vertical polarization. The problem of noise interference with vertical polarization also needs investigating.

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FREQUENCY-MODULATION DISTORTION IN LINEAR NETWORKS

With Special Application to Minimum-Phase-Type Networks

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SUMMARY

The problem of distortion of frequency-modulated waves by 4-terminal linear passive networks has become of increasing importance with the development of multi-channel f.m. communication systems. The paper surveys the somewhat confusing historical background of this subject and after discussion of some essential basic concepts presents a detailed analysis of the problem of small-order distortion. An attempt has been made, using the work of Medhurst,[1] to present the results in a form suitable for computer programming.

Special attention is given to networks which satisfy the minimum-phase condition, because thereby distortion becomes expressible in terms of the amplitude response alone. The results show in a visual yet quantitative manner the error involved in the use of the quasi-stationary solution.

LIST OF PRINCIPAL SYMBOLS

$A(\omega)$ = Amplitude response of network, nepers.
$\phi(\omega)$ = Phase lead of network, rad.
$\psi(\omega)$ = Divergence from linearity of $\phi(\omega)$, rad.
$(H_r)_A$ = Fractional rth harmonic distortion contributed by amplitude response $A(\omega)$.
$[h_r(n)]_A$ = nth sideband amplitude factor associated with $(H_r)_A$.
$(H_r)_\phi$ = Fractional rth harmonic distortion contributed by phase response $\phi(\omega)$.
$[h_r(n)]_\phi$ = nth sideband phase factor associated with $(H_r)_\phi$.
Δ = Angular frequency increment, rad/sec, from ω_0 of origin of linear-amplitude slope.
K, k = Gradient of linear-amplitude slope, nepers/rad sec^{-1} and dB/rad sec^{-1} respectively.
m = Frequency-modulation index.
n = Order of Bessel function $J(m)$.
p = Angular frequency of the signal, rad/sec.
ω = Instantaneous angular frequency, rad/sec.
ω_0 = Angular frequency of the carrier, rad/sec.

(1) INTRODUCTION

It is a well-known fact that intelligence modulated on to a carrier wave becomes distorted in passage through any medium for which the transfer function either is non-linear or has a non-constant frequency response to sinusoidal modulation. Much of the work so far carried out on frequency-type modulation has been in connection with f.m. communication systems with a view to frequency-spectrum economy, and with high-fidelity systems in general, where it is reasonable to assume that the distortion is of small value.

In the paper the intelligence will be assumed to be sinusoidal frequency modulation of a sinusoidal carrier, and the medium will be assumed to be representable by a 4-terminal linear passive network for which the transfer function is known. Even with these restrictions, solution is in general possible only by numerical analysis. Very little theoretical information is available on the extended problem of non-linearity and non-sinusoidal modulation and the effects of noise.

In Section 2 some basic concepts are introduced as being essential to a clear understanding of the subject and of the analysis of small-order distortion which is given in Section 4.

The historical background to the subject is treated in Section 3. Since the early contributions were entirely mathematical and it is only in recent years that serious engineering interest has been aroused, it has been thought appropriate to give a logical presentation based on the two categories of solution defined by Stumpers.[2]

The analysis of small-order distortion given in Section 4 is based on the work of Medhurst[1] and leads to formulation as computer-programming data. Much of the published work has attempted to express the solution in analytical terms. This has led to approximate solutions whose range of validity is ill-defined and to closed-loop solutions of very limited applicability. Numerical analysis is now becoming a practical proposition through the increasing availability of electronic-computer facilities.

(2) BASIC IDEAS

As a prerequisite it is advisable to familiarize oneself with the fundamental theory of modulation. Two excellent tutorial papers are those of Van der Pol[3] (Sections 1 to 3) wherein elementary concepts are made clear, and the more advanced work of Bloch.[4] The mathematical concepts introduced by Bloch in his development of the subject, although probably unfamiliar to the reader, are lucidly explained, and are worth assimilating because of the mental insight provided by the end results. Bloch unifies the general-type modulation (amplitude in combination with phase, or frequency) in a simple functional equation which expresses the modulated-carrier vector in terms of the sum of sideband vectors; the ensuing treatment reveals the linear network in the role of an operator acting upon the sideband vectors.

(2.1) Definition of the Transfer Function of a Network

The network transfer function expresses the effect (output current or voltage) as a ratio of the cause (input current or voltage). The modulus and phase angle of the transfer function define the amplitude and phase response respectively. Thus, referring to Fig. 1, V_1 and V_2 are the input and output terminal voltages, and I_1 and I_2 are the associated terminal currents. For the matched condition of unit terminations, V is the source voltage and I_2' is the associated load current. The transfer function may be defined either as a dimensionless function, V_2/V_1 or I_2/I_1, or as an impedance, V_2/I_1 or V/I_2'. Which definition is used depends of course upon circumstances. For example, if the input and output terminations were pentode valves, the input could be regarded as a constant-current source and the output as an open-circuit termination, and the appropriate function would be V_2/I_1.

Written contributions on papers published without being read at meetings are invited for consideration with a view to publication.

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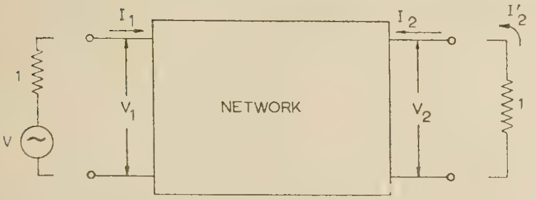

Fig. 1.—Block diagram of a network used to define the amplitude response and phase response.

(2.2) Derivation of Distortion from the Transfer Function for Sinusoidal Input

Consider the input function (voltage or current) to be

$$F_1 \cos\left(\int_0^t \omega dt\right)$$

where $\qquad \omega = \omega_0 + mp \cos pt \quad . \quad . \quad . \quad . \quad . \quad (1)$

is the instantaneous angular frequency.

ω_0 = Angular frequency of the carrier.
p = Angular frequency of the signal.
mp = Angular frequency deviation.

Then the frequency-modulated wave at the input may be written as

$$F_1 \cos(\omega_0 t + m \sin pt)$$

As an aid later to concise formulation of results, let us start from the complex representation and write

$$F_1 \varepsilon^{j(\omega_0 t + m \sin pt)} \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (2)$$

Since the physical problem concerns real quantities, it is implicit in the use of the complex representation that the real part is to be extracted from the final result.

By the standard Fourier-series expansion, the input function

$$F_1 \varepsilon^{j(\omega_0 t + m \sin pt)} = F_1 \varepsilon^{j\omega_0 t} \sum_{n=-\infty}^{\infty} C_n \varepsilon^{jnpt}$$

where
$$C_n = \frac{1}{2\pi}\int_{-\pi}^{\pi} \varepsilon^{jm \sin pt} \varepsilon^{-jnpt} d(pt)$$
$$= \frac{1}{2\pi}\int_{-\pi}^{\pi} \varepsilon^{j(m \sin pt - npt)} d(pt)$$
$$= \mathrm{J}_n(m)$$

Thus
$$F_1 \varepsilon^{j(\omega_0 t + m \sin pt)} = F_1 \varepsilon^{j\omega_0 t} \sum_{n=-\infty}^{\infty} \mathrm{J}_n(m) \varepsilon^{jnpt} \quad . \quad . \quad (3)$$

The derivation leading to eqn. (3) constitutes a concise statement of the Fourier-series expansion of the function given by expression (2). Vaughan[6] gives an interesting informal dissertation on the transform.

The transform of eqn. (3) has the well-known physical interpretation that the infinite series of terms in the right-hand expression are the sideband components of the frequency-modulated wave. It is seen that the sidebands are spaced at intervals equal to the modulating angular frequency p, and that the amplitude of the nth sideband is equal to $\mathrm{J}_n(m)$, the nth order Bessel function of the first kind.

Let $A(\omega)$ nepers and $\phi(\omega)$ radians be the amplitude and phase end of the network at angular frequency ω, relative to their respective values at the carrier angular frequency ω_0. Then the output function (voltage or current) is

$$F_1 \varepsilon^{j\omega_0 t} \sum_{n=-\infty}^{\infty} \mathrm{J}_n(m) \varepsilon^{jnpt} \varepsilon^{A(np) + j\phi(np)} \quad . \quad . \quad . \quad (4)$$

Comparing the output function of expression (4) with the input function of eqn. (3) it is seen that the sideband components have been altered in amplitude and shifted in phase by amounts equal to the amplitude and phase-shift of the network at the respective sideband positions.

For any physically realizable transfer function, exp $[A(\omega) + j\phi(\omega)]$, the output function (4) may be translated into the function

$$F_2 \varepsilon^{j[(\omega_0 t + \theta_0) + K_1 \sin(pt + \theta_1)]} \varepsilon^{j \sum_{n=2}^{\infty} K_n \sin(npt + \theta_n)} \quad . \quad . \quad (5)$$

As expressed in the form of (5), the various coefficients have an immediate physical significance. Thus F_2 is the magnitude of the output function and is, in general, varying at (harmonic) modulation rates; expressed as a fraction of the magnitude F_1 of the input function, it represents amplitude distortion. Frequency-modulation distortion is given by comparing the instantaneous angular frequency ω_{out} at the output with the instantaneous angular frequency at the input. From expression (5),

$$\omega_{out} = \frac{d}{dt}\left[(\omega_0 t + \theta_0) + K_l \sin(pt + \theta_1) + \sum_{n=2}^{\infty} K_n \sin(npt + \theta_n)\right]$$
$$= \omega_0 + K_1 p \cos(pt + \theta_1) + \sum_{n=2}^{\infty} K_n np \cos(npt + \theta_n) \quad . \quad (6)$$

Comparing eqns. (6) and (1) it is seen that the frequency deviation has changed from mp at the input to $K_1 p$ at the output, and this is referred to as linear distortion. Also, harmonic components are present in the output, given by the summation term of eqn. (6).

In general the translation from (4) to (5) must be made by numerical calculation. Artifices have been devised whereby, for special conditions imposed upon the transfer function, the translation may be made analytically. This topic will be pursued further in Section 3.

(3) HISTORICAL BACKGROUND

(3.1) General

All methods of solution are common in that the transfer function of the network is made to operate on the input wave to produce a modified output wave. The methods to be discussed are applicable only to linear networks since they depend upon superposition properties. This is so where the sideband representation of an f.m. wave is used or where the transfer function is expressed in a series form not involving the absolute level of the input wave.

The published methods are classified in two broad divisions which Stumpers[2] has described as Fourier method and asymptotic method. It is under these headings that the aspects of the problem are to be considered.

(3.2) Fourier Method

In the Fourier method the input f.m. wave is expressed in terms of steady-state frequencies known by engineers as the f.m. sidebands of the modulated wave. Each f.m. sideband is weighted by the value of the network transfer function at the respective sideband frequency, and the output f.m. wave is considered to be the vector sum of the weighted sidebands. The concise formulation of the Fourier method by Stumpers in Section 2 of

his work[2] starts from the differential equation expressing the equilibrium of the network. Thus attention is drawn to the transient set up by the driving force at the switching instant. The distortion of the output function occurring over the decay period of the transient is usually of little practical interest. Accordingly, Stumpers rejects the transient part by using the impedance concept to represent the a.c. steady-state response of the network. Toward a better understanding of this, some remarks of Guillemin (pp. 234 and 419–31 of Reference 5) on the topic of network transient analysis would seem relevant.

Medhurst[1] and Assadourian[7] have applied the Fourier method directly to the solution of network problems. In each case a workable form of solution has been obtained by restricting attention to small-order distortion, or more correctly to small divergence of the amplitude and phase response from linearity. This limitation imposes the only validity condition on the results so obtained. In Medhurst's treatment the divergence of the network response from linearity is small but arbitrary. Assadourian considers the special case where the divergence has the form of a sinusoidal ripple.

Under conditions where the divergence of the network response from linearity is not small, the standard method of approach has been to approximate the response by a finite power series or trigonometrical series, and to use sufficient terms to represent the response within a specified accuracy over a given frequency interval. By this artifice it is then possible to change the problem from the operation of the network on an infinite number of sidebands to the operation of the input f.m. wave on the finite number of terms representing the network response. This method would be described more appropriately as a pseudo-Fourier method because, as will be shown below, the final result is closely related to the asymptotic series by consideration of sidebands.

For the power-series representation of the network, Bloch[4] has shown that the final series so obtained is realizable as the asymptotic series of Carson and Fry, terminated at a finite degree. The finite approximation causes an error which may be thought of as arising from sidebands outside the valid range of representation of the network response. For the error to be negligible, a sufficient condition is that the sum, beyond the valid range, of the product (series value at each sideband frequency times the respective sideband magnitude) is negligible.

Frantz,[9] Gold[10] and Collings and Skwirzynski[11] have each followed the pseudo-Fourier method. The approach of Frantz, who has used the Fourier-type trigonometrical series, leads to results of any desired accuracy but seems limited in value by its complexity. Thus the process of computation forms an involved sequence of operations whose end result is not in a convenient form for extracting the harmonic components. Gold, using the power-series representation, has suggested the use of the Legendre or Chebÿshev polynomial to give an approximation to the network response such that the mean-square error is minimum. The output function is expressed in Cartesian form $R + jX$; the polar form $r\varepsilon^{j\phi}$ would be more appropriate since r and $d\phi/dt$ would then express directly the amplitude and frequency distortion. Collings and Skwirzynski have also used the power-series representation, and it is interesting to note that their series result [eqn. (2.1.15) of their paper] is identical with eqn. (5.8) in the paper by Bloch[4] except for symbol notation. In Section 8.4 their analysis is used to obtain the harmonic components of distortion arising from the minimum phase response associated with a linear amplitude slope. The distortion values so obtained follow closely the curves for the quasi-stationary solution discussed in Section 3.3. However, it is shown in Section 4 that the quasi-stationary solution is not a valid one over a wide range of practical conditions. This example should illustrate that in the use of series approximation methods one has to satisfy oneself that the approximation gives a valid representation of the network response over the frequency range of significant sidebands.

Gerlach[12] follows the abstract approach of considering specific network response curves for which the solution by the Fourier method leads to an exact analytic expression. His final conclusion that a linear phase response is a more important design objective than a flat amplitude response would be of especial interest were the supporting evidence more substantial, though in the light of Medhurst's work the converse would be equally true under conditions where the quasi-stationary solution is not valid.

(3.3) Asymptotic Method

In a mathematical study of frequency-modulation theory, Carson and Fry[8] in their eqn. (2.9) obtained an infinite series expansion of the network output function in terms of derivatives of the network transfer characteristic (amplitude and phase response). Stumpers[2] has shown in Section 5 of his article that the Carson and Fry series is asymptotic. In an asymptotic-series expansion of a function, the sum to n terms differs from the value of the function by an amount which, with increase of n, at first decreases, reaches a minimum, and then increases indefinitely as n approaches infinity. It follows from Stumpers's proof that, if the signal angular frequency p is sufficiently small, the network output function may be calculated with great accuracy by taking a suitable number of terms of the series. As p increases, the number of terms before the convergent–divergent point in the series decreases. Accordingly, the possible accuracy falls off until, for large values of p, the series becomes wholly divergent. This process is exemplified in Figs. 6 and 7, where the discrepancy between the function (full-line curves) and the first two terms of the series (dotted-line curves) is shown in terms of the parameter Δ/p. Whittaker and Watson[13] may be consulted for an authoritative reference on page 150 to the definition of an asymptotic expansion.

The reader wishing to follow up Carson and Fry's work will find alternative derivations given in Section 5 of Reference 4, Section 6 of Reference 3 and Section 4 of Reference 2. The first two terms of the Carson and Fry series constitute what is referred to as the quasi-stationary solution, in which the output phase (of which the rate is the output f.m. signal) is taken to be $\phi(\omega_i)$, where ω_i is the instantaneous input frequency, and $\phi(\omega)$ is the steady-state phase characteristic of the system. The reason for the designation is that the results are exact only if the carrier is modulated at an infinitesimal rate, and in fact predict that the output wave may be obtained from the input wave by delaying the instantaneous frequency according to the time-delay response of the network.

Gladwin[14] has used the asymptotic method to obtain a general solution, from which he has deduced series expansions for the cases of large and small values of modulation index. Although Gladwin's results have been referred to by later workers, attention does not appear to have been drawn to certain errors in the text which might be prejudicial to the value of the end results. Thus, rewriting eqn. (3) of Gladwin's paper, the Maclaurin series expansion of $T(\alpha)$ is

$$T(\alpha) = \sum_{n=0}^{\infty} \frac{\alpha^n}{n!} T_n(0) \quad . \quad . \quad . \quad . \quad . \quad (7)$$

where

$$T_n(0) = \frac{d^n}{d\alpha^n} T(\alpha)|_{\alpha=0} \quad . \quad . \quad . \quad . \quad . \quad (8)$$

and

$$\alpha = \frac{\Delta\omega_s - j\dfrac{d}{dt}}{\omega_B}$$

Considering the nth term of eqn. (7), and expanding as a Taylor's series,

$$\frac{\alpha^n}{n!}T_n(0) = \frac{1}{n!}\left[\begin{array}{l}\left(\frac{\Delta\omega_s}{\omega_B}\right)^n + n\left(\frac{\Delta\omega_s}{\omega_B}\right)^{n-1}\left(-j\frac{d}{dt}\right) \\ + \frac{n(n-1)}{2!}\left(\frac{\Delta\omega_s}{\omega_B}\right)^{n-2}\left(-j\frac{d}{dt}\right)^2 + \ldots\end{array}\right]T_n(0) \quad . \quad . \quad . \quad . \quad (9)$$

Assuming it to be permissible to substitute eqn. (9) in eqn. (7) we have

$$T(\alpha) = \sum_{n=0}^{\infty}\frac{1}{n!}\left(\frac{\Delta\omega_s}{\omega_B}\right)^n T_n(0)$$

$$+ \sum_{n=0}^{\infty}\frac{1}{(n-1)!}\left(\frac{\Delta\omega_s}{\omega_B}\right)^{n-1}\left(-j\frac{d}{dt}\right)T_n(0) + \ldots \quad . \quad (10)$$

The first term of eqn. (10) has the form of eqn. (7) and agrees with the first term of eqn. (4) of Gladwin's paper. However, there is no agreement between succeeding terms. In the above the question of convergency has not been considered, though such an examination would be necessary to attach meaning to results obtained from the series expansion.

Hupert[15] has used the first two terms of the series in eqn. (2.9) of Carson and Fry[8], namely the quasi-stationary solution, and by further approximation has related the orders of harmonic components of distortion to the respective orders of derivatives of the network phase response. Hupert's systematized solution is to be commended, but, because it is restricted in validity by the boundaries of the quasi-stationary solution, it may easily lead to fallacious results. For example, the quasi-stationary solution is not applicable to ripples in the amplitude and phase response adjacent to the region of frequency deviation. Because of the asymptotic nature of the Carson and Fry series, Hupert's extension of the analysis to include the third term of the series leads to results of greater accuracy only within the range of validity of the quasi-stationary solution.

(4) ANALYSIS OF SMALL-ORDER DISTORTION

(4.1) General

For the condition of small distortion, Medhurst[1] has shown that each harmonic component of the frequency-modulation distortion consists of two independent expressions, one a function of the amplitude response and the other a function of the phase response, and that these add in quadrature. By analogy with the physical interpretation of expression (4), the first expression may be thought of as arising from amplitude change of the sideband vectors, and the second as arising from phase shift of the sideband vectors. The final results of Medhurst are so indispensable to the analysis which follows that a fuller derivation than that of Medhurst is given in Section 8.2. The manner of approach differs slightly in order to show more clearly the artifices whereby the general solution of expression (4) has been translated analytically into the usable form of expression (5). The separate contributions to distortion made by the amplitude and by the phase response are considered in Sections 4.2 and 4.3 respectively.

In Section 8.1 some notes are given on the minimum-phase condition, since in Section 4.4 this condition is imposed to enable the phase response to be expressed in terms of the amplitude response. From the above it follows that small-order distortion of a frequency-modulated wave by a minimum-phase-type network is completely determinate in terms of the amplitude response only. The practical value of this conclusion must be stressed. As is shown in Section 8.2, linear variation of phase with frequency produces a time delay of the signal but does not give rise to distortion. Consequently, it is the divergence of the phase response from linearity which must be measured. Not only will the divergence be small for small-order distortion, but experimental techniques for precise measurement of phase shift are very much less straightforward than those for precise measurement of amplitude.

The treatment of Section 4.4 hinges upon certain linear superposition properties whereby the distortion is expressible in an effective form for design purposes. First, the amplitude response is broken down into a number of linear amplitude slopes. The minimum-phase response associated with the amplitude response is then given by the algebraic sum of the phase contributions of the separate linear amplitude slopes. This is a consequence of the linear relation existing in the minimum-phase transform between amplitude and minimum phase. Finally, the algebraic sum of the distortion contributions of the separate amplitude slopes is equal to the distortion arising from the overall minimum-phase response. This is a consequence of the linear relation existing between phase shift and the dependent distortion. Thus the method enables the relative importance of the various portions of the amplitude response to be readily assessed, as also the effect of small changes in amplitude.

It is, of course, preferable to compute the phase harmonic distortions directly from the phase response when this is available, but it is possible, nevertheless, to estimate these distortions from the amplitude response alone when the network is known to be of minimum-phase type.

(4.2) Amplitude of Sideband Vectors

From eqns. (26) and (27) the expressions for the contribution to distortion made by the amplitude response $A(\omega)$, expresed as a percentage, are

$$\left.\begin{array}{l}(H_2)_A = \dfrac{200}{m}\displaystyle\sum_{n=1}^{\infty} J_n(J_{n-2} - J_{n+2})[A(np) - A(-np)] \text{ per cent} \\ \text{for the second-harmonic component, and} \\ (H_3)_A = \dfrac{300}{m}\displaystyle\sum_{n=1}^{\infty} J_n(J_{n-3} - J_{n+3})[A(np) + A(-np)] \text{ per cent}\end{array}\right\} (11)$$

for the third-harmonic component.

The argument of the Bessel functions J_n is the modulation index m.

Let us define amplitude harmonic-distortion factors

$$\left.\begin{array}{l}[h_2(n)]_A = \dfrac{1}{8\cdot686}\left[\dfrac{200}{m}J_n(J_{n-2} - J_{n+2})\right] \text{ per cent} \\ [h_3(n)]_A = \dfrac{1}{8\cdot686}\left[\dfrac{300}{m}J_n(J_{n-3} - J_{n+3})\right] \text{ per cent}\end{array}\right\} \quad . \quad (12)$$

where the factor 8·686 is the conversion factor from nepers to decibels.

Then

$$\left.\begin{array}{l}(H_2)_A = 8\cdot686\displaystyle\sum_{n=1}^{\infty}[h_2(n)]_A[A(np) - A(-np)] \text{ per cent} \\ (H_3)_A = 8\cdot686\displaystyle\sum_{n=1}^{\infty}[h_3(n)]_A[A(np) + A(-np)] \text{ per cent}\end{array}\right\} \quad . \quad (13)$$

In eqn. (13) the units of the amplitude $A(np)$ are nepers, but may be converted to decibels by omission of the factor 8·686. The reason for the designation 'distortion factor' to parameters $[h_2(n)]_A$ and $[h_3(n)]_A$ follows from eqn. (13), which, in words, states that the resultant harmonic distortion is given by summation of the amplitude (relative to the amplitude at carrier

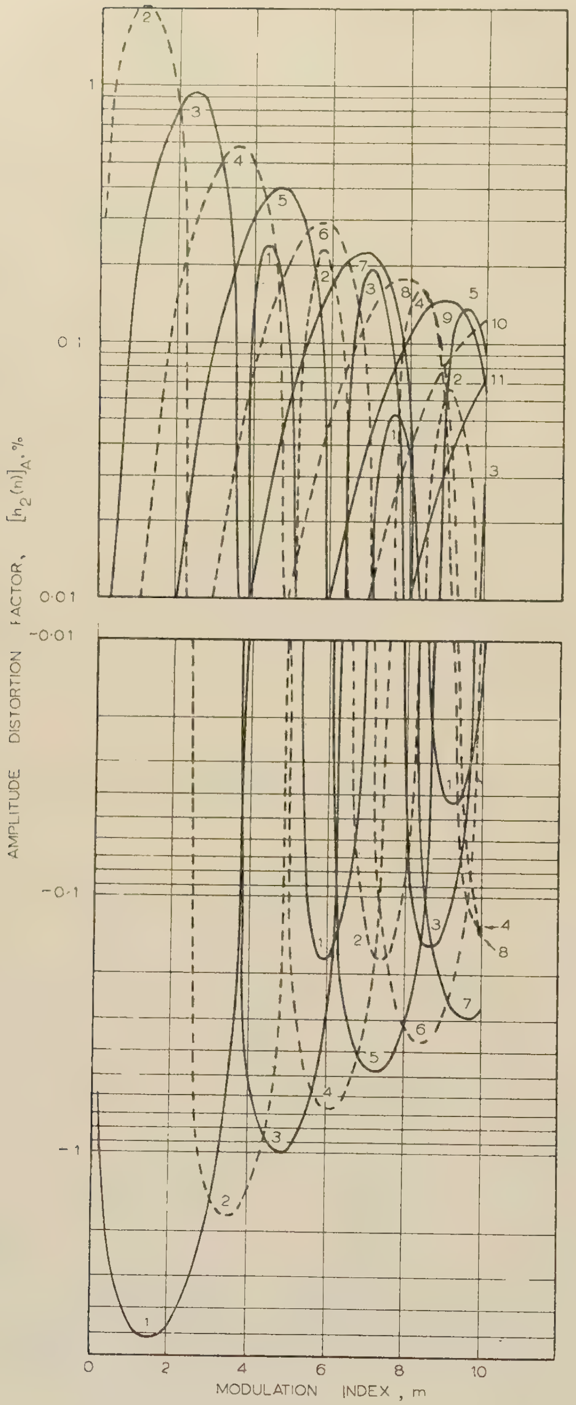

Fig. 2.—Sideband curves for the second-harmonic amplitude-distortion factor.

——— Odd-order sidebands.
– – – Even-order sidebands.

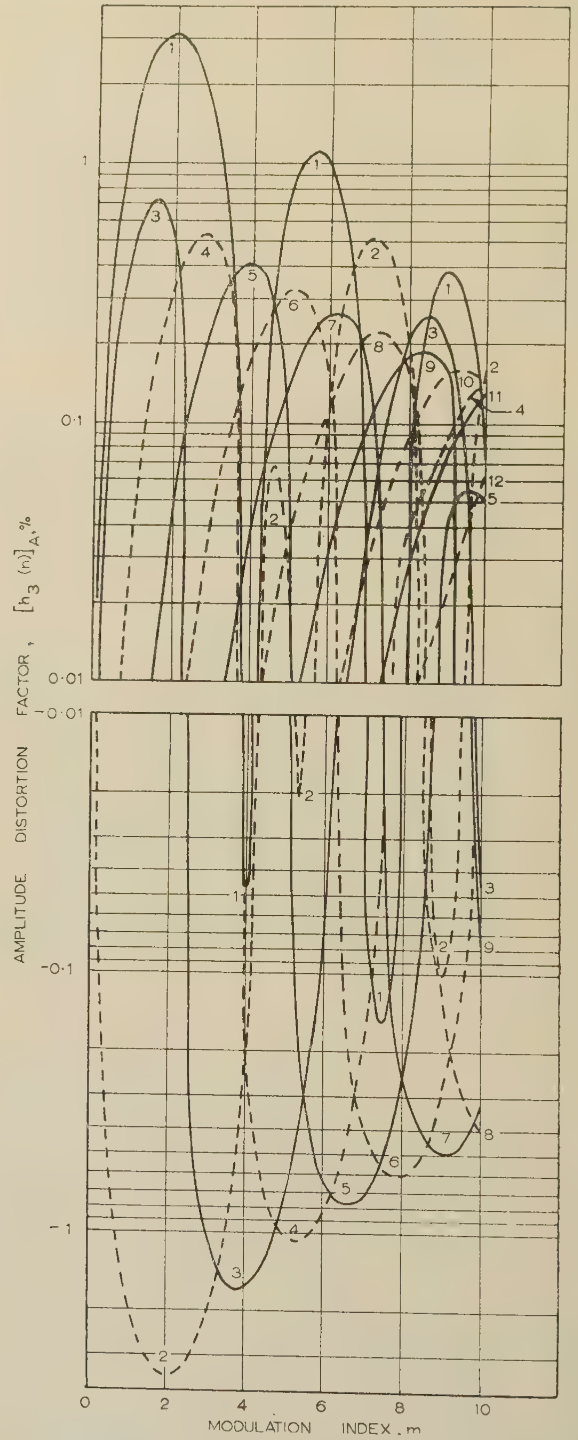

Fig. 3.—Sideband curves for the third-harmonic amplitude-distortion factor.

——— Odd-order sidebands.
– – – Even-order sidebands.

frequency) at each sideband frequency, and weighted by the respective distortion factor.

The distortion factors of eqn. (12) are plotted in Figs. 2 and 3 as functions of the modulation index m with the sideband number n as parameter. To save space the sideband curves are superimposed on the one graph, but for the purpose of illustration a typical sideband curve, $n = 1$ for the second harmonic distortion, has been redrawn in Fig. 4. The oscillatory nature of the curve is at once seen, as also the rapid rate of change of the distortion factor for values of m in the region of cross-over points. These observations, since they apply equally well to the other sideband curves, bring out the critical dependence of distortion upon the modulation index m. Such dependence was not evident from the work of Medhurst. From inspection of

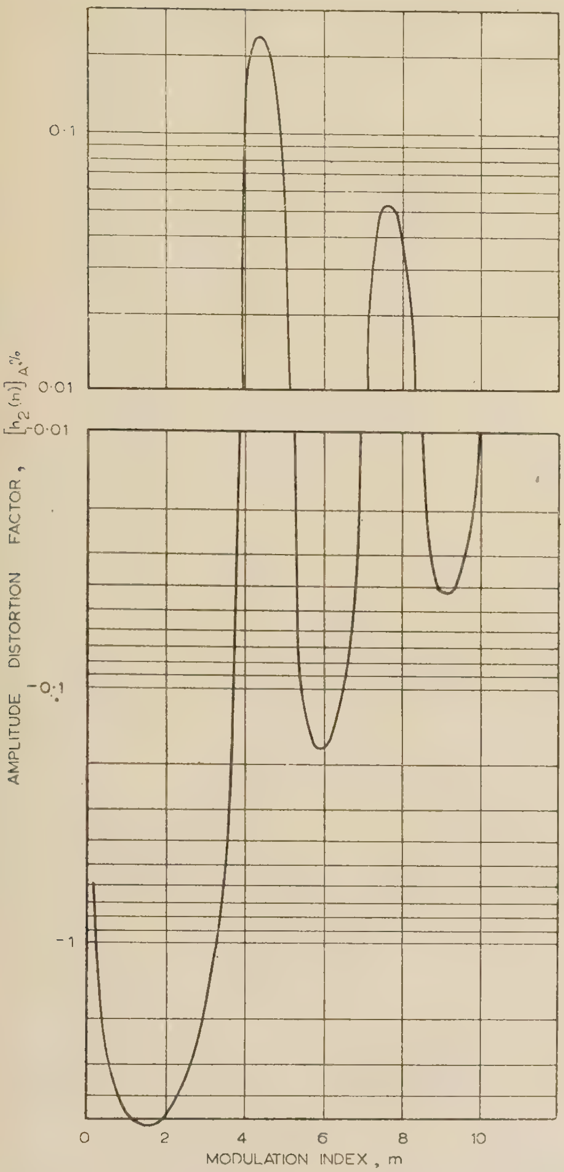

Fig. 4.—Typical sideband curve ($n = 1$) for the second-harmonic amplitude-distortion factor.

Figs. 2 and 3, large changes in distortion may be expected for the lower values of m, although at the higher values variations will be smoothed out by the increasing number of significant sidebands. The practical aspects are important because the modulation index m is dependent upon the linear distortion of the network [see eqn. (6)] and, further, in voice transmission the modulation index is not definable except in statistical terms.

The sign of the distortion factor (as plotted in Figs. 2 and 3) must, in the case of the second harmonic, be reversed for the sidebands below the carrier frequency. This may be expected in order that the second harmonic distortion be zero for an amplitude response having arithmetical symmetry.

(4.3) Phase Shift of Sideband Vectors

From eqns. (26) and (27) of Section 8.2, the expressions for the contribution to distortion made by the phase divergence $\psi(np)$, expressed as a percentage, are

$$\left.\begin{aligned}(H_2)_\phi &= -\frac{200}{m}\sum_{n=1}^{\infty} J_n(J_{n-2}+J_{n+2})\\&\qquad[\psi(np)+\psi(-np)]\text{ per cent}\\&\text{for the second-harmonic component and}\\(H_3)_\phi &= -\frac{300}{m}\sum_{n=1}^{\infty} J_n(J_{n-3}+J_{n+3})\\&\qquad[\psi(np)-\psi(-np)]\text{ per cent}\end{aligned}\right\} \quad . \; (14)$$

for the third-harmonic component.

In like manner to eqn. (12), let us define phase harmonic factors

$$\left.\begin{aligned}[h_2(n)]_\phi &= -\frac{200}{m}J_n(J_{n-2}+J_{n+2})\text{ per cent}\\ [h_3(n)]_\phi &= -\frac{300}{m}J_n(J_{n-3}+J_{n+3})\text{ per cent}\end{aligned}\right\} \quad . \quad (15)$$

Then

$$\left.\begin{aligned}(H_2)_\phi &= \sum_{n=1}^{\infty}[h_2(n)]_\phi[\psi(np)+\psi(-np)]\text{ per cent}\\(H_3)_\phi &= \sum_{n=1}^{\infty}[h_3(n)]_\phi[\psi(np)-\psi(-np)]\text{ per cent}\end{aligned}\right\} \quad . \quad (16)$$

(4.4) Minimum-Phase Condition

Consider the amplitude response to be approximated by a number of straight lines connecting successive points along the response curve, the curve being thereby broken down into segments which represent linear steps in the amplitude value. A linear step such as AB shown in Fig. 5(a) may be treated as the

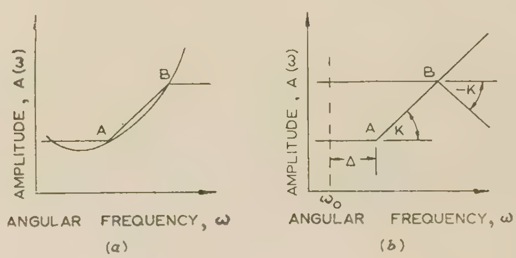

Fig. 5.—Synthesis of amplitude response from linear-amplitude slopes.

(a) Linear amplitude step.
(b) Representation of linear-amplitude step by two linear-amplitude slopes.

sum of two linear amplitude slopes each of which terminates at infinity. This is shown in Fig. 5(b) from which it is seen that the purpose of the second slope is to cancel the rate of change of amplitude beyond point B. It follows that any amplitude response may be synthesized from a number of linear amplitude slopes. Attention is therefore restricted to consideration of an arbitrary linear amplitude slope of gradient k nepers/rad sec^{-1}, located at angular frequency increment Δ from the carrier angular frequency ω_0.

The minimum phase shift $\phi(\omega)$ associated with a linear amplitude slope of parameters (Δ, K) is, from the minimum-phase transform on page 335 of Reference 16,

$$\phi(\omega) = \frac{K}{\pi}\int_{\omega_0+\Delta}^{\infty}\log\left|\frac{\Omega+\omega}{\Omega-\omega}\right|d\Omega \quad . \quad . \quad . \quad . \quad (17)$$

Integrating eqn. (13),

$$\phi(\omega) = \frac{K}{\pi}\Big[(\Omega+\omega)\log(\Omega+\omega) - (\Omega-\omega)\log|\Omega-\omega|\Big]_{\Omega=\omega_0+\Delta}^{\Omega=\infty}$$

$$= \frac{K}{\pi}\Big[2\omega\log\Omega + (\Omega+\omega)\log\left(1+\frac{\omega}{\Omega}\right) - (\Omega-\omega)\log\left|1-\frac{\omega}{\Omega}\right|\Big]_{\Omega=\omega_0+\Delta}^{\Omega=\infty}$$

Let $x = \omega - \omega_0$.

$$\text{Then}\quad \phi(x) = \frac{K}{\pi}\left[\begin{array}{c} -2(x+\omega_0)\log(\omega_0+\Delta) \\ -(2\omega_0+x+\Delta)\log\left(1+\dfrac{\omega_0+x}{\omega_0+\Delta}\right) \\ +(\Delta-x)\log\left|1-\dfrac{\omega_0+x}{\omega_0+\Delta}\right| \end{array}\right]$$

For the purpose of evaluating distortion we are interested only in the divergence $\psi(x)$ of $\phi(x)$ from linearity; as demonstrated in Section 8.2 $\psi(x)$ may be obtained by subtracting from $\phi(x)$ the first-order approximation in x to $\phi(x)$.

Thereby

$$\psi(x) = \frac{K}{\pi}\left[\begin{array}{c} 2x + (\Delta - x)\log\left|1-\dfrac{x}{\Delta}\right| \\ -(2\omega_0+x+\Delta)\log\left(1+\dfrac{x}{2\omega_0+\Delta}\right) \end{array}\right]$$

At sideband positions $x = np$, so that

$$\psi(np) = \frac{npK}{\pi}\left[\begin{array}{c} 2+\left(\dfrac{\Delta}{np}-1\right)\log\left|1-\dfrac{np}{\Delta}\right| \\ -\left(\dfrac{2\omega_0+\Delta}{np}+1\right)\log\left(1+\dfrac{np}{2\omega_0+\Delta}\right) \end{array}\right] \quad (18)$$

From eqn. (18) it is seen that, of the two parameters (Δ, K) which define the linear amplitude slope, the parameter K appears only as a scale factor. For convenience, K will be replaced by the frequency-normalized parameter pK; further, the units of pK will be converted from nepers to decibels by the relation

$$8{\cdot}686pK \text{ nepers} = pk \text{ decibels},$$

to give

$$\frac{\psi(np)}{pk} = \frac{n}{8{\cdot}686\pi}\left[\begin{array}{c} 2+\left(\dfrac{\Delta}{np}-1\right)\log\left|1-\dfrac{np}{\Delta}\right| \\ -\left(\dfrac{2\omega_0+\Delta}{np}+1\right)\log\left(1+\dfrac{np}{2\omega_0+\Delta}\right) \end{array}\right] \quad (19)$$

For each value of n, $\psi(np)/pk$ is seen to be a function of Δ/p and $(2\omega_0+\Delta)/p$. Under narrow-band conditions, eqn. (19) simplifies to the approximate form

$$\frac{\psi(np)}{pk} \simeq \frac{n}{8{\cdot}686\pi}\left[1+\left(\frac{\Delta}{np}-1\right)\log\left|1-\frac{np}{\Delta}\right|\right]\left(\frac{2\omega_0+\Delta}{p} \gg 1\right) \quad (20)$$

(4.5) Procedure of Numerical Calculation

The reader must convince himself by further reading, but may for the moment accept, that frequency-modulation distortion problems are soluble by mathematical means only by lengthy computation. So-called simple methods of solution will be found either to be simple in appearance only, or to have achieved their simplicity at the expense of dependability of the end results. This state of affairs is not peculiar to frequency-modulation distortion, of course, but exists in many practical problems. The increasing availability to industry of digital-computer facilities is enabling the mathematical approach to a problem to be of value even where it leads to lengthy computation. The major task then becomes the setting-up of the problem as a programme of instructions to be applied to the computer. To this end the distortion problem is here reduced to a set of addition-multiplication instructions which is repeated for each sideband (amplitude and phase components of distortion). For the minimum-phase condition the phase component of distortion may be obtained from a similar set of instructions which is repeated for each linear amplitude slope.

(4.5.1) Amplitude and Phase Components of Distortion.

The harmonic components of distortion are given by eqns. (13) and (16). The setting-up of the problem requires the tabulation of the amplitude value at each sideband number relative to the value at carrier frequency, and of the divergence of the phase from linearity at each sideband number. Each amplitude or phase value is weighted by its associated distortion factor at the respective sideband number, and then the weighted values are summed to give the distortion corresponding to the harmonic number of the distortion factor. Thus the procedure requires tabulation of the harmonic distortion factor. The graphs in Figs. 2 and 3 should be of help in deciding the range of tabulation for the amplitude component, and it is suggested that tabulation at intervals in modulation index m of 0·2, to an accuracy of three significant figures, should be adequate for most purposes.

(4.5.2) Minimum-Phase Condition: Phase Component of Distortion.

On substituting eqn. (19) in eqn. (16), it is possible to tabulate $(H_2)_\phi/pk$ and $(H_3)_\phi/pk$ as functions of Δ/p and $(2\omega_0+\Delta)/p$

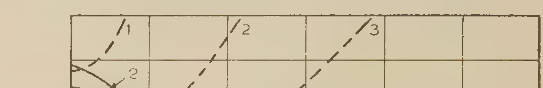

Fig. 6.—Second-harmonic distortion arising from the minimum-phase response associated with a linear-amplitude slope.

——— Method of Medhurst.
– – – Quasi-stationary solution.
Numbers denote values of Δ/p.

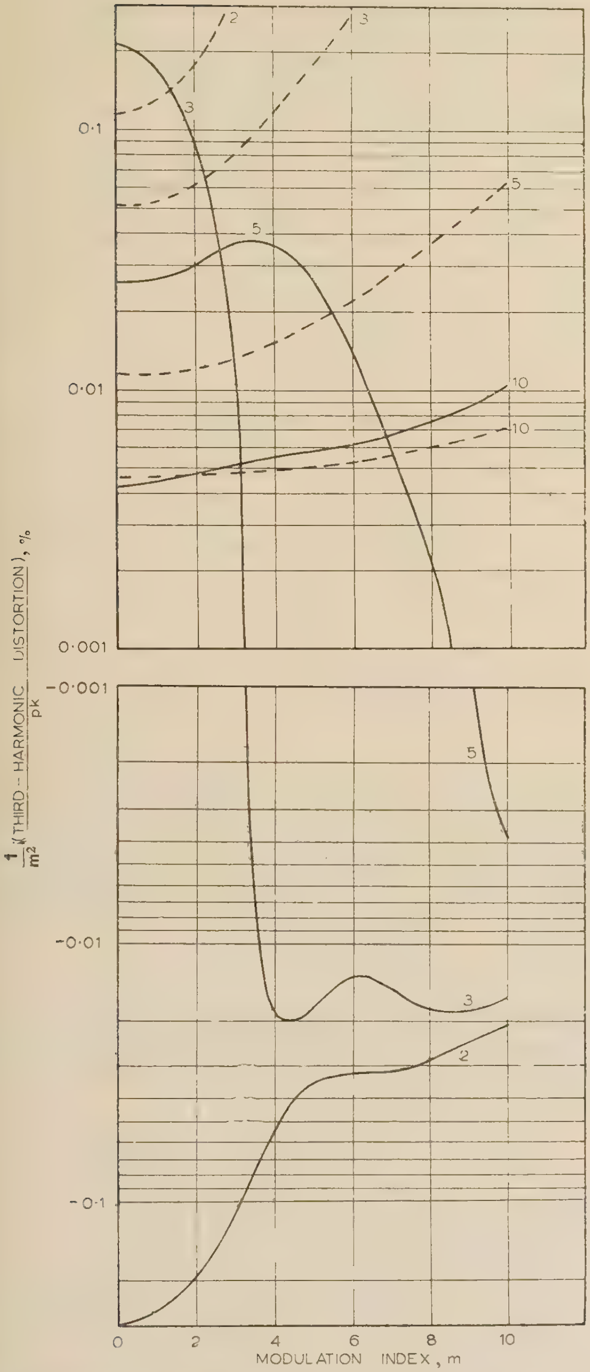

Fig. 7.—Third-harmonic distortion arising from the minimum-phase response associated with a linear-amplitude slope.

——— Method of Medhurst.
– – – Quasi-stationary solution.
Numbers denote values of Δ/p.

only. The process of tabulation is laborious. However, once it is completed, the setting-up of the problem requires only the tabulation of the (Δ, k) parameter of the linear amplitude slopes used to approximate the amplitude response. The type of problem to which this work has been applied has satisfied the narrow-band condition expressed by eqn. (20), and therefore $(H_2)_\phi/pk$ and $(H_3)_\phi/p$ have so far been tabulated only as a function of Δ/p. The results are plotted in Figs. 6 and 7 (full-line curves). It is of interest to compare the quasi-stationary predictions (dotted curves), whose derivation is given in Section 8.3.

In the initial design it may be of value to postulate an equivalent circuit for the network, and hence deduce a theoretical amplitude and phase response. Thereby an analytic expression for ψ may be substituted directly in eqn. (16) rather than indirectly by computation from the amplitude response [eqn. (19)]. For example, a network of widespread practical interest is the one formed by the valve-coupled interstage circuits of an intermediate-frequency amplifier; the interstage circuits are usually single- or double-tuned circuits. Theoretical derivation of the amplitude and phase response for such a network has been treated quite exhaustively in the literature. The attention of the reader is drawn to the analysis by Rudd[17] for the double-tuned circuit, since the choice of parameters leads to a most convenient solution [eqns. (31) to (37) of Rudd's paper] for extraction of ψ, the divergence of the phase response from linearity.

(5) CONCLUSIONS

The distortion of a frequency-modulated wave by a 4-terminal passive network appears to have no analytic solution. Where workers have made approximations to obtain an analytic solution their criteria for range of validity are in general expressed in terms too vague for reliable usage. This criticism applies to the detailed analysis of the problem of small-order distortion given in Section 4, in that a probable error arising from the assumption of small-order distortion cannot be applied to the end results.

The analysis in Section 4 was intended for applications where the tolerable distortion was less than about -60 dB, under which condition the results obtained are reliable. For the method to be of practical value computer facilities are required. Although the results are in a form suitable for computer programming, tables of instructions need to be computed, and to this end some of the results are given in the form of graphs.

The form of treatment given in the paper is based upon the linear superposition properties of linear networks. Since linear networks produce amplitude-modulation distortion of a frequency-modulated wave as well as frequency-modulation distortion, amplitude limiters must be inserted to remove the amplitude modulation. In the literature it is often tacitly assumed that the amplitude-limiting process does not of itself introduce frequency-modulation distortion. This is not true in general, and there is need for investigation of the distortion produced by non-linear networks, of which amplitude limiters are an example.

(6) ACKNOWLEDGMENTS

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(8) APPENDICES

(8.1) Restrictions Imposed Upon a Network by the Minimum-Phase Condition

In the following, appeal will be made to the concept of zeros and poles of the network transfer function and their location in the complex frequency plane. The reader will find a helpful exposition on these matters given on pages 286 *et seq.* of Reference 5.

For a signal input of the form ε^{st}, the transfer function $F(s)$ of the network may be written as the ratio of two polynomials $P(s)$ and $Q(s)$ which, factorized in terms of their roots, gives

$$F(s) = \text{constant} \times \frac{(s-s_1)(s-s_3)\ldots(s-s_{2n-1})}{(s-s_2)(s-s_4)\ldots(s-s_{2n})}$$

The roots $s_1, s_3, \ldots$ and the roots $s_2, s_4, \ldots$ are respectively the zeros and poles of the transfer function. For physical realizability of the network, poles cannot occur in the right-hand half of the complex-frequency plane (r.h. plane). Zeros may, however, occur in the r.h. plane, but Bode[16] has shown (p. 236) that these zeros may be transformed to poles in the left-hand plane by introducing all-pass networks. The residual network, having zeros and poles in the left-hand plane only, is defined to be the minimum phase-shift network. The reason for this designation is that any further rearrangement of the zeros and poles must result in a change in the amplitude response.

In general, a network cannot be specified to be of minimum-phase type until the location of its zeros and poles is known. However, a passive ladder network is always that type.

(8.2) Harmonic Components of Distortion by the Method of Medhurst

Re-stating expression (4), the output function is

$$G_2 = F_1\varepsilon^{j\omega_0 t}\sum_{n=-\infty}^{\infty} J_n(m)\varepsilon^{jnpt}\varepsilon^{A(np)+j\phi(np)} \quad . \quad . \quad (21)$$

For the special case where the amplitude response is linear, i.e.

$$A(np) = np \times \text{constant } c_1,$$

and the phase response is linear, i.e.

$$\phi(np) = np \times \text{constant } c_2,$$

$$G_2 = F_1\varepsilon^{j\omega_0 t}\sum_{n=-\infty}^{\infty} J_n(m)\varepsilon^{jnp(t+c_2-jc_1)}$$

which by the transform of eqn. (3)

$$= F_1\varepsilon^{j[\omega_0 t + m\sin p(t+c_2-jc_1)]}$$

$$= F_1\varepsilon^{m\cos p(t+c_2)\sinh pc_1}\varepsilon^{j[\omega_0 t+m\sin p(t+c_2)\cosh pc_1]} \quad . \quad (22)$$

Comparing eqn. (22) with eqn. (2), it is seen that a combination of linear amplitude and phase characteristics causes linear frequency-modulation distortion but does not introduce harmonic distortion. Medhurst quoted this result as justification for assuming that, for small values of $A(np)$, $\phi(np)$ might be replaced by a function

$$\psi(np) = \phi(np) - npc_2$$

which represents the divergence of the phase response from linearity. $\psi(np)$ is small for small distortion.

For the purpose of estimating harmonic distortion, the output function may be written as

$$G_2' = F_1\varepsilon^{j\omega_0 t}\sum_{n=-\infty}^{\infty} J_n(m)\varepsilon^{jnpt}\varepsilon^{A(np)+j\psi(np)}$$

where both amplitude $A(np)$ and phase lead $\psi(np)$ are considered to be small.

To first-order approximation,

$$G_2' \simeq F_1\varepsilon^{j\omega_0 t}\sum_{n=-\infty}^{\infty} J_n(m)\varepsilon^{jnpt}[1 + A(np) + j\psi(np)]$$

$$= F_1\varepsilon^{j\omega_0 t}\left\{\varepsilon^{jm\sin pt} + \sum_{n=-\infty}^{\infty} J_n(m)\varepsilon^{jnpt}[A(np) + j\psi(np)]\right\}$$

which by the transform of eqn. (3)

$$= F_1\varepsilon^{j\omega_0 t}(\varepsilon^{jm\sin pt} + R + jX) \quad . \quad . \quad . \quad (23)$$

where $$R = \sum_{n=-\infty}^{\infty} J_n(m)[A(np)\cos npt - \psi(np)\sin npt]$$

$$X = \sum_{n=-\infty}^{\infty} J_n(m)[A(np)\sin npt + \psi(np)\cos npt]$$

If eqn. (23) can be translated into the form

$$G_2' = F_1\varepsilon^{j\omega_0 t}\varepsilon^{jm\sin pt}\varepsilon^{S+jY} \quad . \quad . \quad . \quad . \quad (24)$$

then by comparison with eqns. (5), (6), the harmonic distortion is given by dY/dt.

Since R and X are small, so also will be S and Y. Therefore, to first-order approximation,

$$G_2' \simeq F_1\varepsilon^{j\omega_0 t}[\varepsilon^{jm\sin pt} + \varepsilon^{jm\sin pt}(S + jY)]$$

By comparison with eqn. (23),

$$R = S\cos(m\sin pt) - Y\sin(m\sin pt)$$

$$X = S\sin(m\sin pt) + Y\cos(m\sin pt)$$

Therefore $Y = X\cos(m\sin pt) - R\sin(m\sin pt)$.

In the complex representation,

$$Y = (X + jR)\varepsilon^{jm\sin pt}$$

$$= \sum_{n=-\infty}^{\infty} J_n(m)[jA(np) + \psi(np)]\varepsilon^{-jnpt}\varepsilon^{jm\sin pt}$$

by substitution of values for X and R.

By the transform of eqn. (3),

$$Y = \sum_{n=-\infty}^{\infty} J_n(m)[jA(np) + \psi(np)]\varepsilon^{-jnpt}\sum_{r=-\infty}^{\infty} J_r(m)\varepsilon^{jrpt}$$

$$= j\sum_{n=-\infty}^{\infty}\sum_{r=-\infty}^{\infty} J_n(m)A(np)J_r(m)\varepsilon^{-j(n-r)pt}$$

$$+ \sum_{n=-\infty}^{\infty}\sum_{r=-\infty}^{\infty} J_n(m)\psi(np)J_r(m)\varepsilon^{-j(n-r)pt}$$

$$\frac{dY}{dt} = p\sum_{n=-\infty}^{\infty}\sum_{r=-\infty}^{\infty} J_n(m)A(np)(n-r)J_r(m)\varepsilon^{-j(n-r)pt}$$

$$- jp\sum_{n=-\infty}^{\infty}\sum_{r=-\infty}^{\infty} J_n(m)\psi(np)(n-r)J_r(m)\varepsilon^{-j(n-r)pt} \quad . \quad (25)$$

As already interpreted from eqn. (24), dY/dt is the harmonic distortion. From eqn. (25) the second-harmonic component, expressed as a fraction of the frequency deviation mp, is

$$H_2 = \frac{1}{m}\sum_{n=-\infty}^{\infty} J_nA(np)(2J_{n-2}\varepsilon^{-j2pt} - 2J_{n+2}\varepsilon^{j2pt})$$

$$- \frac{j}{m}\sum_{n=-\infty}^{\infty} J_n\psi(np)(-2J_{n+2}\varepsilon^{j2pt} + 2J_{n-2}\varepsilon^{-j2pt})$$

where all the Bessel functions have argument m.

Extracting the real part of the expression,

$$H_2 = \frac{2}{m}\sum_{n=-\infty}^{\infty} J_nA(np)(J_{n-2} - J_{n+2})\cos 2pt$$

$$+ \frac{2}{m}\sum_{n=-\infty}^{\infty} J_n\psi(np)(-J_{n+2} - J_{n-2})\sin 2pt$$

Using the relation that for a positive integer n,

$$J_{-n} = (-)^n J_n$$

$$H_2 = \frac{2}{m}\sum_{n=1}^{\infty} J_n(J_{n-2} - J_{n+2})[A(np) - A(-np)]\cos 2pt$$

$$- \frac{2}{m}\sum_{n=1}^{\infty} J_n(J_{n-2} + J_{n+2})[\psi(np) + \psi(-np)]\sin 2pt \quad . \quad . \quad . \quad . \quad (26)$$

Likewise the third-harmonic component, expressed as a fraction of the frequency deviation mp, is

$$H_3 = \frac{3}{m}\sum_{n=1}^{\infty} J_n(J_{n-3} - J_{n+3})[A(np) + A(-np)]\cos 3pt$$

$$- \frac{3}{m}\sum_{n=1}^{\infty} J_n(J_{n-3} + J_{n+3})[\psi(np) - \psi(-np)]\sin 3pt \quad . \quad . \quad . \quad . \quad (27)$$

Eqns. (26) and (27) correspond to eqns. (13) and (14) of Medhurst.

(8.3) Distortion Arising from the Minimum-Phase Response associated with a Linear Amplitude Slope, according to the Quasi-Stationary Solution

Below is given the derivation of the quasi-stationary results plotted in Figs. 6 and 7.

According to the quasi-stationary solution, the frequency-modulation distortion F, expressed as a fraction of the angular frequency deviation mp, is

$$F = \frac{1}{mp}\frac{d}{dt}\phi(\omega) = \frac{1}{mp}\frac{d}{d\omega}\phi(\omega)\frac{d}{dt}(\omega_0 + mp\cos pt)$$

$$= -p\frac{d}{d\omega}\phi(\omega)\sin pt$$

To evaluate the distortion, the divergence $\psi(\omega - \omega_0)$ from linearity [eqn. (20), where $\omega - \omega_0 = np$] may replace $\phi(\omega)$. Therefore

$$F = -p\frac{d}{dx}\psi(x)\sin pt$$

$$= -\frac{pk}{8\cdot 686\pi}\frac{d}{dx}\left[\frac{x}{\Delta} + \left(1 - \frac{x}{\Delta}\right)\log\left|1 - \frac{x}{\Delta}\right|\right]\sin pt$$

where

$$x = \omega - \omega_0$$

$$= mp\cos pt$$

For the condition $x/\Delta < 1$, the logarithmic term may be expanded into a series, whence

$$F = -\frac{pk}{8\cdot 686\pi}\left[\frac{x}{\Delta} + \frac{1}{2}\left(\frac{x}{\Delta}\right)^2 + \frac{1}{3}\left(\frac{x}{\Delta}\right)^3 + \ldots\right]\sin pt$$

Extracting the second- and third-harmonic components, F_2 and F_3 respectively, of F by Fourier analysis, and expressing as a percentage,

$$F_2 = -\frac{100pk}{8\cdot 686\times 2\pi}\left[\frac{mp}{\Delta} + \frac{1}{6}\left(\frac{mp}{\Delta}\right)^3 + \frac{1}{16}\left(\frac{mp}{\Delta}\right)^5 + \ldots\right]\text{per cent}$$

$$F_3 = -\frac{100pk}{8\cdot 686\times 8\pi}\left[\left(\frac{mp}{\Delta}\right)^2 + \frac{3}{8}\left(\frac{mp}{\Delta}\right)^4 + \frac{3}{16}\left(\frac{mp}{\Delta}\right)^6 + \ldots\right]\text{per cent} \quad . \quad . \quad . \quad . \quad (28)$$

For the graphs in Figs. 6 and 7, eqns. (28) have been arranged in the form

$$\frac{1}{m}\left(\frac{F_2}{pk}\right) = -1\cdot 832\frac{p}{\Delta}\left[1 + \frac{1}{6}\left(\frac{mp}{\Delta}\right)^2 + \frac{1}{16}\left(\frac{mp}{\Delta}\right)^4 + \ldots\right]\text{per cent}$$

$$\frac{1}{m^2}\left(\frac{F_3}{pk}\right) = -0\cdot 458\left(\frac{p}{\Delta}\right)$$

$$\left[1 + \frac{3}{8}\left(\frac{mp}{\Delta}\right)^2 + \frac{3}{16}\left(\frac{mp}{\Delta}\right)^4 + \ldots\right]\text{per cent} \quad . \quad . \quad . \quad . \quad (29)$$

Thus in the range where the quasi-stationary solution may be expected to hold, the scaled values of distortion [eqn. (29)] should be invariant with change in modulation index m.

(8.4) Distortion Arising from the Minimum-Phase Response associated with a Linear Amplitude Slope, according to Collings and Skwirzynski

Collings and Skwirzynski[11] assume that the amplitude and phase response of the network may each be expanded as a power series in $x = \omega - \omega_0$, and restrict analysis to the sixth power in x. In the present instance the amplitude response is constant (unity), and the phase response is given by eqn. (20).

For the condition $x/\Delta < 1$, where $x = \omega - \omega_0$ replaces np, eqn. (20) may be expanded as the series

$$\psi(x) = \frac{k\Delta}{8\cdot 686\pi}\left[\begin{array}{l}\frac{1}{2}\left(\frac{x}{\Delta}\right)^2 + \frac{1}{6}\left(\frac{x}{\Delta}\right)^3 + \frac{1}{12}\left(\frac{x}{\Delta}\right)^4 \\ + \frac{1}{20}\left(\frac{x}{\Delta}\right)^5 + \frac{1}{30}\left(\frac{x}{\Delta}\right)^6 + \ldots\end{array}\right]$$

$$= \sum_{s=2}^{\infty} a_s\left(\frac{x}{\Delta}\right)^s \quad \ldots \ldots \quad (30)$$

Eqn. (30) defines the coefficients a_s of eqn. (2.1.7) of Collings and Skwirzynski's paper.

Following their method of analysis to obtain harmonic coefficients H_c, H_s (pp. 122, 123 of their paper) as a function of a_s,

$$H_{c2} = -\,a_2a_3(3m^2 + 14)mp^5 \quad \ldots \ldots \quad (31)$$

$$H_{s2} = -\,a_2mp^2 - a_4(m^2 + 7)mp^4$$
$$-\,a_6(\tfrac{15}{16}m^4 + 20m^2 + 31)mp^6 + a_2^3(\tfrac{4}{3}m^2 + \tfrac{14}{3})mp^6 \quad . \quad (32)$$

$$H_{c3} = -\,\tfrac{3}{2}a_2^2m^2p^4 - a_2a_4(\tfrac{9}{2}m^2 + \tfrac{123}{2})m^2p^6 - a_3^2(\tfrac{81}{32}m^2 + \tfrac{63}{2})m^2p^6 \quad \ldots \quad (33)$$

$$H_{s3} = -\,\tfrac{3}{4}a_3m^2p^3 - a_5(\tfrac{15}{16}m^2 + \tfrac{75}{4})m^2p^5 \quad \ldots \ldots \quad (34)$$

The above results agree with the tabulated results of Collings and Skwirzynski, except for eqn. (33), which they give as

$$H_{c3} = -\,\tfrac{3}{2}a_2^2m^2p^4 - a_2a_4(\tfrac{9}{2}m^2 + \tfrac{123}{2})m^2p^6$$
$$-\,a_3^2(\tfrac{81}{32}m^2 + \tfrac{63}{2})m^2p^6 + a_6(\tfrac{135}{2})m^2p^6 \quad . \quad (35)$$

The discrepancy between eqns. (33) and (35) is too small to affect the remainder of the argument.

From eqn. (30), coefficients a_s are seen to vary directly as k, the gradient of the linear amplitude slope. The value of k may always be made sufficiently small that only first-order terms in k need to be considered. Thus for k arbitrarily small, eqns. (31) to (34) reduce to

$$H_{c2} = 0$$
$$H_{s2} = -\,a_2mp^2 - a_4(m^2 + 7)mp^4 - a_6(\tfrac{15}{16}m^4 + 20m^2 + 31)mp^6$$
$$H_{c3} = 0$$
$$H_{s3} = -\,\tfrac{3}{4}a_3m^2p^3 - a_5(\tfrac{15}{16}m^2 + \tfrac{75}{4})m^2p^5$$

Substituting values for a_s from eqn. (30), and expressing the distortion as a percentage,

$$H_{s2} = -\frac{100pk}{8\cdot 686 \times 2\pi}\left[\begin{array}{l}\frac{mp}{\Delta} + \frac{1}{6}\left(1 + \frac{7}{m^2}\right)\left(\frac{mp}{\Delta}\right)^3 \\ + \frac{1}{16}\left(1 + \frac{64}{3m^2} + \frac{496}{15m^4}\right)\left(\frac{mp}{\Delta}\right)^5\end{array}\right] \quad (36)$$

$$H_{s3} = -\frac{100pk}{8\cdot 686 \times 8\pi}\left[\left(\frac{mp}{\Delta}\right)^2 + \frac{3}{8}\left(1 + \frac{20}{m^2}\right)\left(\frac{mp}{\Delta}\right)^4\right] \quad . \quad (37)$$

Comparing eqns. (36) and (37) with the quasi-stationary solution given by eqn. (28), it is seen that the solution according to Collings and Skwirzynski follows closely the quasi-stationary solution. But it must be remembered that, in the derivation of eqn. (28) and of eqns. (36) and (37), the phase response was expressed in series form on the understanding that $x/\Delta < 1$. In the sideband representation of the f.m. wave this condition is that $np/\Delta < 1$ and, for validity, must hold up to the largest value n of a significant sideband.

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THE EROSION OF ELECTRICAL CONTACTS BY THE NORMAL ARC

By W. B. ITTNER, III, Ph.D., and H. B. ULSH, M.A.

(*The paper was first received* 21*st March, and in revised form* 21*st September*, 1956.)

SUMMARY

The cathode material transfer under the action of the 'normal arc' as been measured for a number of elements and alloys. In all cases uxiliary circuits were used to record the accumulative arc duration nd the voltage/current characteristics of the arc, from which the total rc energy as well as the total charge passed in the arc could be ccurately calculated. Within the experimental errors the material ransfer from the cathode was found to be directly proportional to he total charge passed in the arc—a relationship first proposed by R. Holm. It is shown that, for most practical purposes, the cathode normal arc' transfer can be calculated with a reasonable accuracy ccording to the formula given by Llewellyn Jones.

LIST OF SYMBOLS

V_m = Minimum arc voltage.
V_i = Contact ionization potential.
ϕ = Contact work function.
I_m = Minimum arc current.
$I_{R\,max}$ = Maximum current in arc in resistive circuit.
$I_{L\,max}$ = Maximum current in arc in inductive circuit.
I_0 = Closed-circuit current.
L = Circuit inductance.
R = Circuit resistance.
v = Contact opening velocity.
w = Material transfer.
Q = Charge passed in arc.
k = Normal-arc matter-transfer coefficient.
α, β, γ = Arbitrary constants.
λ = Contact thermal conductivity.
T = Contact boiling temperature.
ρ = Contact density.
c = Contact specific heat.
A = Contact atomic weight.

(1) INTRODUCTION

Erosion under the action of the so-called 'normal arc' is one of the determining factors in the practical operating life of lectrical contacts which are required to interrupt currents of he order of several amperes to inductive or resistive load devices. While a number of authors[1,2,3] have discussed quantitatively he physical mechanisms of the normal arc and the factors which might be expected to influence matter transfer, only a small mount of reliable experimental data on the magnitude of arc ransfer is available. The lack of data collected under controlled onditions has led the authors to reinvestigate the magnitude of he arc transfer of a number of common elements and alloys nder conditions where the arc duration and energy could be ccurately measured by means of simple auxiliary circuits.

(2) FUNDAMENTALS

An outline of the various transfer processes which accompany n arc between contacts has been given by Llewellyn Jones.[2] It is convenient to discuss the material erosion produced by the arc in terms of the voltage/current characteristics of the arc with gap separation. Such discussion leads to a recognition of three more or less distinct types of arc; namely the short arc, the normal arc and the high-power arc.

(*a*) The short arc, which has been the subject of a number of recent investigations[4–8] is associated with contact separations of the order of 10^{-4} cm, and is characterized in most cases by a predominant vaporization of material from the anode.* This vaporization is produced by the high-speed electrons which cross the gap without appreciable energy loss and bombard the anode directly.

(*b*) At greater contact separations the formation of the 'normal arc' plasma tends to 'cushion' the anode from direct electron bombardment. The bulk of the arc energy is dissipated in the plasma and at the surface of the cathode, which is bombarded directly by positive ions accelerated through the cathode fall of potential. In the normal arc, therefore, the bulk of evaporation occurs at the cathode surface.

(*c*) For extremely long or high-power arcs, the high current density and the onset of an appreciable anode drop combine to produce a marked increase in the anode evaporation. For high-power arcs, therefore, appreciable evaporation will occur at both anode and cathode.

The material loss or transfer of material from one contact to the other which takes place under the action of the arc may occur by a number of processes. The evaporated material may, for example, diffuse thermally from the site of vaporization and subsequently condense on a cooler region of either contact. Such a process is referred to as 'thermal transfer'. In addition, a fraction of the evaporized contact material, which, in general, has a lower ionization potential than the gas which is present in the arc, will become ionized, and these metallic ions may be carried across the contact gap under the action of the applied electric field. This process is referred to as 'ionic transfer'. The transfer of material or the material loss or gain of either contact may also be effected by electrostatic forces which act on the molten-metal contact surfaces, by mechanical vibrations or impacts which would tend to shear off small deposits from the contact faces, and by chemical reactions which take place between the contact materials and the surrounding atmosphere, resulting in the formation of oxides, hydrides and nitrides, or the decomposition and deposition of environmental contaminants on the contact surfaces.

Because of the variety of processes which may occur it is possible to envisage situations in which both contacts lose or gain material, as well as situations where one contact loses material, a fraction of which is gained by the other. In spite of the general complexity of the overall problem, controlled measurements of the arc transfer carried out in a region where one type of arc transfer clearly predominates have been found to be

Written contributions on papers published without being read at meetings are vi…ed for consideration with a view to publication.
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* L. H. Germer and W. S. Boyle have recently shown that short arcs are of two fundamental types, those characterized by a predominant anode loss of material (anode arcs), and those in which the loss of material is confined to the cathode (cathode arcs). (Paper No. D–3, 8th Annual Gaseous Electronics Conference, Schenectady New York, 20th October, 1955.)

reasonably consistent and may be used to evaluate contact materials with regard to their relative merit for practical use.

The paper describes measurements of the 'normal arc' transfer which takes place in resistive and inductive 50-volt circuits, interrupting currents of 1·5–5 amp. The arc duration under these conditions ranged from 1 to 11 millisec, corresponding to contact gaps of 10^{-2} to 0·3 cm. The principal mode of transfer consisted of material loss at the cathode, a fraction of which was generally deposited on the anode.

(3) THE NORMAL ARC

The normal arc may be illustrated by means of well-known voltage/current characteristics for gap separations as shown hypothetically in Fig. 1, where the voltage/current characteristics

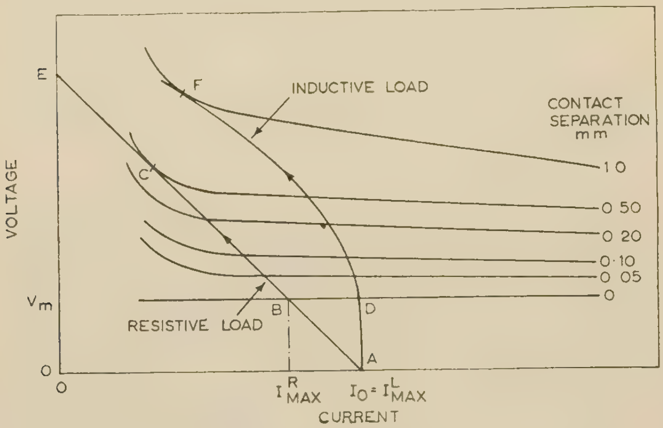

Fig. 1.—Typical voltage/current arc characteristics for a number of contact separations.

AB is a dynamic load line for a resistive circuit.
AF is a typical load line for an inductive circuit.

of the arc have been plotted for a number of distances of separation of the contacts.[1] The arc is characterized by a so-called minimum arc voltage V_m, which is approximately equal to the sum of the work function, ϕ, and the ionization potential, V_i, of the contact material, and by a so-called minimum arc current I_m. It has recently been shown by Atalla[9] that the minimum arc current is not constant but is a function of the maximum current in the arc and is influenced appreciably by the surface condition of the contacts; i.e. oxidized or chemically activated surfaces exhibit a minimum arc current which may be several orders of magnitude lower than the minimum arc current of the clean metal surface. The voltage/current curves must therefore be considered as only rather rough approximations of the actual arc characteristics, which may change not only with time but also from one arc to the next, depending on the surface state of the contacts.

Neglecting for the present the changing nature of the voltage/current characteristics it is convenient to use these curves to discuss the observable characteristics of the arc which occur when a pair of contacts is used to interrupt a current. In the case of an essentially resistive load the arc is constrained to a load-line curve such as AE in Fig. 1, where A corresponds to the situation in which the closed contacts carry the full load current I_0 under a negligible voltage drop. When the contacts are opened the voltage rises to the minimum arc voltage V_m, and an arc is first initiated at a current, $I_{R\,max}$ (point B). As the contacts begin to separate the arc voltage rises and the arc current decreases until it is finally extinguished at a point where the load line first becomes tangential to the voltage and current/distance curves (point C). Figs. 2(*a*) and 2(*b*) show typical voltage and current/time curves for a single arc in a resistive circuit.

Fig. 2.—Typical voltage/time and current/time curves for an arc in a resistive circuit.

(*a*) Voltage/time.
(*b*) Current/time.

In the case of an inductive circuit the situation is considerably more complex. For a very large inductance, i.e. large value of L/R, the effect of the inductance is to attempt to maintain the current flowing through the contacts prior to rupture. At break, therefore, the contact voltage rises to the minimum arc voltage with only a negligible decrease in the circuit current, and the arc is initiated at a current, $I_{L\,max}$ (D in Fig. 1), very nearly equal to the closed-circuit current I_0. As the contacts begin to separate, the arc voltage increases and the circuit current decreases in a manner indicated roughly by the curve DF. The actual shape of the curve will be determined by the value of L/R of the load and the opening velocity v of the contacts. In general, the greater the ratio of L/R to v, the steeper is the slope of this curve. In time the curve DF will begin to slope in such a manner as to join the static curve AE, with the arc being extinguished at some point such as F. Typical voltage/time and current/time curves for an arc in an inductive circuit are shown in Figs. 3(*a*) and 3(*b*). Because the contact surfaces are changing with time, however, the points at which the arc is extinguished (C and F in Fig. 1) will vary from operation to operation, and it is difficult to utilize such characteristic curves to obtain more than a rough approximation to the arc characteristics.

(4) EXPERIMENTAL METHOD

Thus it is almost impossible to describe analytically the arc characteristics in terms of the physical parameters of the circuit. This is particularly true since quantities such as the arc duration, the charge passed in the arc, and the arc energy depend in a rather complex manner on the circuit time-constant, L/R, the opening velocity, v, of the contacts, and the mathematical expressions for the voltage/current gap-separation curves of the arc, which are themselves only approximate. On the other hand, it is a relatively simple matter to measure the characteristics of the arc by means of standard laboratory apparatus.

Such measurements, carried out with an oscilloscope and a time-interval meter, have shown that the arc characteristics with a given circuit not only vary statistically from cycle to cycle,

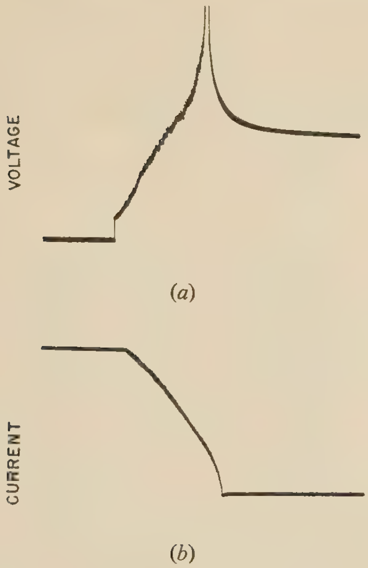

Fig. 3.—Typical voltage/time and current/time curves for an arc in an inductive circuit.

(a) Voltage/time.
(b) Current/time.

but may also vary systematically with time; i.e. in most cases the arc duration increases systematically as clean contacts are run repeatedly and gradually become activated or oxidized. Such an increase in the arc duration is shown, for example, in Fig. 4, where the average arc duration (over 2 000 successive operations in a typical test run) is plotted as a function of the total number of arcs at a pair of platinum contacts. Fig. 4 is, moreover,

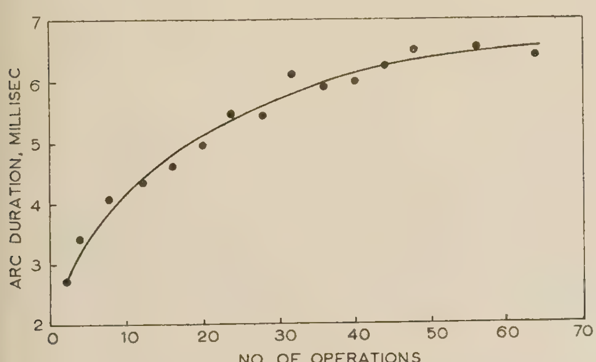

Fig. 4.—A typical plot of the average arc duration as a function of the total number of operations of the contacts.

characteristic of the results obtained with most materials. It is obvious, therefore, that the contact transfer per operation, in a given circuit, will not be a linear function of the number of operations unless the contacts have been run in to the extent that the arc duration has reached some equilibrium value. While most materials eventually reach an equilibrium (as shown in Fig. 4), the number of operations required depends on the circuit parameters, the contact material and the average arc energy.

In the work to be described, all arc durations were measured by means of simple auxiliary circuits which could be used to measure either individual arc durations or accumulative arc durations (for the purpose of obtaining an average arc length). By this means, statistical and systematic variations in the arc duration or arc energy could be measured, making possible a relatively precise determination of total arc energy and total charge passed in the arc. It was found that, on the whole, the cathode transfer could be described quite accurately in terms of the total charge passed during the arc, which in most practical cases is almost directly proportional to the arc energy. On the other hand, while the cathode loss per coulomb of arc current was found to be essentially constant in the range investigated, the anode gain ranged from 3 to 80% of the cathode loss, and was in numerous cases quite erratic. (In general, the fractional anode gain decreased with increasing arc duration and current.) The measurements of more fundamental importance, therefore, appear to be the cathode loss per coulomb of arc current, which are the measurements reported in the paper. The charge passed by the arc and its energy were computed from the arc characteristics, as measured oscilloscopically, and the average arc duration as measured with a time-interval recorder. Weight measurements of the cathode loss and anode gain were carried out on a commercial microbalance with a reproducibility of about $\pm 1\,\mu$g.

A schematic of the circuit used in the investigation is shown in Fig. 5. The break contacts, S_2, which were circular discs

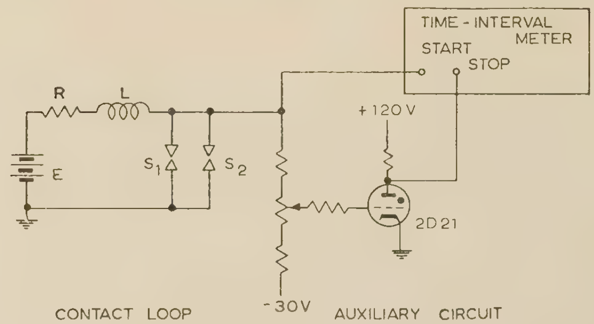

Fig. 5.—Circuit used to measure the average arc duration.

4·75 mm in diameter, were mounted in a mechanically operated circuit-breaker which was driven by one of a number of cams. The cam profile, as well as the speed of the driving motor, could be varied to change the opening velocity of the contacts in the range 10–25 cm/sec. To eliminate transfer due to a short arc or bounce on making the circuit, a separate set of contacts S_1 was used to establish the circuit current. Contact timing could be adjusted quite simply by an appropriate choice of driving cams. The load consisted of various resistors and inductors mounted in a ventilated housing where they were air cooled to prevent excessive heating. By varying the resistive and inductive components of the load and the opening velocity of the contacts, the arc duration, the arc energy and the charge passed in the arc could be varied over a reasonably wide range.

The time-interval meter was adjusted to start with the opening of the break contacts and to stop with the contact-voltage jump at the termination of the arc.

(5) EXPERIMENTAL RESULTS

The experimental results for normal arc transfer (cathode loss) substantiate the relationship proposed by Holm[1] that the material transfer, w, is proportional to the total charge, Q, passed in the arc,

i.e. $$w = kQ$$

where k is defined as the normal-arc matter-transfer coefficient, conveniently expressed in microgrammes per coulomb. This relationship was found to be valid within the experimental errors of 15–20% over the entire current range 1·5–5·0 amp, independent of the circuit load and the current/voltage charac-

Table 1

Material	Composition by weight	Arc transfer coefficient (cathode loss)	
	%	$g \times 10^{-6}$/C	$cm^3 \times 10^{-6}$/C
Palladium		19·4	1·59
Platinum		8·7	0·41
Silver		3·6	0·34
Tungsten		1·1	0·06
Palladium–copper	50–50	20·5	1·94
Platinum–iridium	90–10	17·6	0·82
Platinum–iridium	85–15	16·5	0·77
Platinum–iridium	80–20	14·5	0·67
Palladium–ruthenium	95–5	14·1	1·15
Platinum–ruthenium	89–11	13·2	0·65
Platinum–iridium	65–35	11·6	0·54
Silver–platinum	97–3	5·8	0·55
Silver–palladium	97–3	4·7	0·45
Silver–nickel–palladium–copper	60–23–12–5	3·5	0·25

teristics of the arc. Table 1 contains the average measured values of the matter-transfer coefficient k for a number of elements and alloys.

It is of interest to note that, in most but not all cases, the alloying of materials appears to result in a transfer coefficient which lies somewhere between the coefficients of the constituents themselves. In general, both the arc initiation voltage and the minimum arc current of all the alloys investigated appeared to be characteristic of the element with the smaller value, while the plasma gradient of the arc (as measured from the voltage/time slope) appeared to vary with the alloy composition. For example, all the platinum–iridium alloys exhibited the minimum arc voltage and minimum arc current of iridium, with the plasma gradient increasing slightly with increasing iridium content.

The matter-transfer coefficients in col. 1 of Table 1 are expressed in terms of weight loss (i.e. microgrammes per coulomb) corresponding to the experimentally measured quantity. This can be converted to cubic centimetres per coulomb by dividing by the density of the contact material, which has been done in obtaining the volume loss of material listed in column 2.

Some discretion must be exercised, however, in interpreting the results of Table 1. Differences, for example, in the minimum arc voltages, the minimum arc currents and the plasma gradients of the various materials result in different total arc currents being passed by the different materials under a particular set of conditions. As an example, the matter-transfer coefficients of palladium and palladium–copper are, within the experimental errors, essentially equal. However, owing to the lower minimum arc potential and minimum arcing current of the alloy, the normal arc erosion of palladium–copper is greater than that of palladium in any given circuit. Table 2 contains, as an example, the average arc durations and average arc currents for a number of the materials in Table 1, when these materials were run in a typical test circuit. Multiplication of the average charge passed per arc, as obtained from Table 2, col. 2, by the corresponding transfer coefficients obtained from Table 1, gives the average arc transfer per operation for the various materials in this particular instance (cols. 3 and 4).

(6) DISCUSSION

It is generally agreed that material transfer under the action of the normal arc is principally a vaporization process in which the heat available for vaporization at the cathode is equal to the total heat supplied to the cathode by the arc less the heat which is conducted and radiated from the cathode spot. Exact calculations of these quantities are difficult owing to inexactnesses in the thermal conductivity at high temperatures and a lack of knowledge of the arc current densities and the fraction of the current carried by positive ions. In addition, it is known that the cathode spot may move about the cathode surface so that thermal equilibrium is not always established. It is, however, possible to carry out rough calculations of the total energy available for vaporization at the cathode, leading to formulae which indicate the expected magnitude of the cathode transfer for various materials. Llewellyn Jones,[10] for example, has obtained a relationship which was found to predict reasonably well the transfer coefficients of high-voltage spark-plug electrodes and which should be valid for normal-arc transfer. This equation in a slightly modified and simplified form is

$$k = \frac{\alpha - \beta T^4 - \gamma \lambda T}{\rho\left(c + \dfrac{21}{A}\right)T}$$

where k, the matter transfer coefficient in cubic centimetres per coulomb, is expressed in terms of three arbitrary constants, α, β, and γ; the thermal conductivity, λ, in joules per second-centimetres per degree centigrade; the boiling temperature, T, in degrees centigrade; the density, ρ, in grammes per cubic centimetre; the specific heat, c, in joules per gramme-degree centigrade; and the atomic weight, A, of the contact material. The three arbitrary constants, α, β, and λ, are related respectively to the fraction of the total arc energy available for evaporation at the cathode, the fraction of the energy lost from the cathode by radiation and the fraction of the energy lost from the cathode by conduction. The denominator is an approximation to the energy required to raise the cathode material to the boiling point and evaporate it.

Using $\alpha = 2{\cdot}4 \times 10^{-2}$, $\beta = 2{\cdot}04 \times 10^{-17}$ and $\gamma = 2{\cdot}52 \times 10^{-6}$, a reasonable fit is obtained between the measured and

Table 2

Material	Average arc duration	Average charge passed in arcs	Arc transfer per operation (cathode loss)	
	millisec	mC	$g \times 10^{-9}$	$cm^3 \times 10^{-9}$
Palladium	3·49	5·57	110	9·0
Platinum	2·43	4·32	38	1·8
Silver	3·06	5·05	18	1·7
Palladium–copper (50%–50%)	3·81	6·17	127	12·0
Platinum–iridium (90%–10%)	2·85	4·43	78	3·6
Palladium–ruthenium (95%–5%)	3·78	6·62	94	7·4
Platinum–ruthenium (89%–11%)	3·73	4·63	61	3·0
Platinum–iridium (65%–35%)	2·72	4·21	48	2·2
Silver–platinum (97%–3%)	3·76	6·21	36	3·4
Silver–palladium (97%–3%)	3·85	6·17	29	2·8
Silver–nickel–palladium–copper (60%–23%–12%–5%)	3·05	4·95	17	1·2

Table 3

Material	ρ	T	λ	c	A	Transfer coefficient (cathode loss)	
						Measured	Calculated
	g/cm³	°C	joules/sec-cm per deg C	joules per g-deg C		cm³ × 10⁻⁶/C	cm³ × 10⁻⁶/C
Ag	10·5	1950	4·2	0·23	108	0·33	0·34
Al	2·7	2057	2·1	0·89	27	1·4	10·0*
Au	19·3	2600	2·9	0·13	197	0·34	1·1*
Cu	8·9	2330	3·6	0·38	64	0·12	1·0*
Fe	7·8	3000	0·59	0·46	56	0·92	2·0*
Ir	22·4	4800	0·58	0·13	193	0·25	
Mo	10·0	3700	1·2	0·26	96	0·51	
Ni	8·7	2900	0·54	0·45	59	0·91	
Pd	12·2	2200	0·70	0·29	107	1·51	1·59
Pt	21·4	4300	0·70	0·13	195	0·43	0·41
Pt–10% Ir ..	21·6	4400	0·31	0·13	194	0·63	0·81
Pt–20% Ir ..	21·7	4500	0·18	0·13	194	0·60	0·67
Pt–35% Ir ..	21·8	4600	0·20	0·13	194	0·54	0·54
Rh	12·4	3880	0·88	0·24	103	0·42	0·5*
Sn	7·2	2660	0·65	0·22	119	3·10	
W	19·3	5900	1·5	0·14	184		0·06
Zn	7·1	907	1·1	0·39	65	5·0	5·4*

* Measured by R. Holm (see Reference 1, page 386).

calculated transfer coefficients, as shown in Table 3, which, in addition to data collected by the authors, also contains some previous data of Holm.[1] The agreement in most cases is quite good, considering the approximations which have been made in arriving at the above formula, although there are some notable exceptions, e.g. aluminium and copper. While it is possible to obtain a more exact formula by considering, for example, the temperature dependence of λ, and the size of the cathode spot, in general the exact numerical values are unknown and invalidate the usefulness of such attempts. Considering the data as a whole, Llewellyn Jones's formula is certainly useful in obtaining a first approximation to the expected cathode arc transfer of most materials. A number of common materials have been included in Table 3.

Holm[11] has also given a similar equation for the energy balance at the cathode spot. Holm has taken into account the fractional positive ion current and the current density of the arc. His results differ from those of Llewellyn Jones in that the constant γ in Llewellyn Jones's equation should, according to Holm, contain a factor of the form a/I, where a is the radius of the cathode spot and I is the total arc current. According to the calculations of Holm, therefore, the constant γ should decrease with increasing current, as a smaller fraction of the energy available for vaporization is conducted away from the cathode spot. Such an effect should give a transfer coefficient which increases with increasing current. No such trend was observed in the current range reported in this work, although at higher currents there appeared to be a definite increase in the transfer coefficient. It is re-emphasized, therefore, that the results in Table 3 are for the current range of 1–5 amp.

Llewellyn Jones's formula may be used to explain qualitatively the behaviour of the transfer of the various platinum–iridium alloys in Table 3. The effect of alloying iridium with platinum is first to increase the transfer, which eventually begins to decrease as the percentage of iridium is increased. The increase in the transfer of the alloys containing a small amount of iridium is obviously a result of a decrease in the thermal conductivity of these alloys as compared with platinum.

The contacts were examined microscopically and, in some cases, with the aid of X-ray diffraction apparatus, in an attempt to obtain some information concerning the arc current-density and the formation of by-products on the contact surfaces.

Microscopic examination of the cathode surface following a single arc revealed a number of pits which were formed as the arc moved about on the surface. Such a movement is evidence that a thermal equilibrium is not maintained during the life of the arc and is another argument against rigorous calculations of the cathode transfer. It was not possible to obtain an exact value for the cathode current density owing to the fact that the current itself was decreasing throughout the life of the arc. It was possible, however, to set a lower limit of 10^6 amp/cm^2, with the actual current density being somewhat greater. This is in good agreement with the value of approximately 10^7 amp/cm^2 which may be inferred from Holms's equation for the cathode energy balance[11] by assuming a fractional ion current equal to 10% of the total arc current. After the contacts had been subjected to prolonged arcing the cathode surface appeared as a highly polished metallic area surrounded by a region where discoloration indicated the presence of a tarnish. The tarnish appeared greater with the silver and silver-alloy materials than with the platinum–palladium, and platinum and palladium alloys.

The anode was in all cases covered with a blackened deposit which appeared in greater quantity on the palladium and platinum material than on the silver materials. An analysis of this residue showed it to be composed largely of carbonaceous material containing some finely divided metal particles and some traces of copper oxide. The carbonaceous material was obviously deposited from the surrounding atmosphere, while the copper oxide was apparently formed from copper impurities in the contact materials. As has been pointed out, these deposits, whether carbonaceous or oxide, had the effect of lowering the minimum arc current and prolonging the arc.

In the course of this work, attempts were made to measure the erosion coefficients of several refractory-type commercially available contact materials. In all cases these contacts failed after a few thousand operations owing to the formation of an insulating flux-like material on the contact surfaces. While the composition of this film is unknown, it appears to be formed of a flux or wetting material used in the manufacturing process.

(7) CONCLUSIONS

In the current range 1–5 amp all the materials investigated showed a normal arc transfer which was proportional to the

charge passed in the arc. The agreement between the measured normal-arc matter-transfer coefficients and those computed from the modified formula of Llewellyn Jones is sufficiently good to permit evaluation of the normal-arc transfer coefficients by the use of this formula. Accordingly, the desired material properties for minimizing the normal-arc erosion are a high boiling temperature, a high thermal conductivity and a high density.

The arc transfer coefficients are to be used as a guide for the selection of contact materials. However, with regard to their ability to withstand arc erosion, it is necessary to know the average arc current per operation in order to determine the optimum material (as regards erosion) for a given application. While no simple method is available to determine the arc length in a given application, in general the higher the minimum arc voltage, the higher is the minimum arc current, and the higher is the plasma gradient; the arc length is shorter and the total charge passed in the arc is smaller. In practice, an oscilloscope can be used to measure the arc duration and current/time curves from which the total arc current can be calculated. If the volume of transferred material which constitutes failure can be estimated, the operating life of the contacts with regard to material transfer can be estimated with a fair degree of accuracy. If the boiling temperature, thermal conductivity and density of a particular alloy are known, Llewellyn Jones's formula provides a relatively easy means of estimating the expected transfer in the current range 1–5 amp.

Obviously more information on the energy dissipation at the electrodes, the current densities of the arc, and the fraction of the current carried by positive ions, is necessary to extend these results to apply to arc erosion at high currents.

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NORTH-EASTERN RADIO AND MEASUREMENT GROUP: CHAIRMAN'S ADDRESS

By A. E. TWYCROSS, Associate Member.

'THE MERGING OF RADIO AND LAND-LINE TECHNIQUES'

(Abstract *of Address delivered at* Newcastle upon Tyne, 15*th October*, 1956.)

The demand for communication channels continues to increase, and various schemes have been developed to obtain more channels requiring less external line plant.

Overhead construction was the standard for many years, an additional circuit sometimes being obtained by superposing.

Loaded underground air-spaced paper-core cables, with 2-wire (bothway) amplifiers, permitted many more channels to be obtained, with considerable economy in size of cable conductors, and superposing could be fully exploited because of the greater stability of cable pairs. By changing the lay of the cable from multiple-twin to star-quad formation, one-third more wires could be accommodated in a given cable diameter. The limitations of the 2-wire amplifier were overcome, and much larger overall gains secured, by using unidirectional amplifiers and separate pairs for the 'go' and 'return' directions. Separate pairs involved large multi-pair cables, and separate amplifiers necessitated fairly large buildings, spaced about 50 miles apart.

Modulating a carrier frequency by an audio frequency and extracting the sidebands, as done by the radio engineer, together with the development of negative-feedback principles, enabled 12-channel, and later 24-channel, carrier systems to be used with unloaded cables, the amplifier spacings being reduced to 22 and 12 miles respectively. This enabled a 24-pair cable to accommodate 288 or 576 channels. The coaxial-cable system, developed simultaneously, enabled 600 channels to be carried by one tube, but the amplifier spacing had to be reduced to 6·1 miles.

The trend in design is therefore to increase the number of channels per cable or tube at the expense of providing more and closer spaced intermediate amplifiers, and this would seem to set the limit of the land-line system.

It is at this stage that the potentialities of the s.h.f. radio-link system becomes attractive, as with the broader bandwidths easily obtained, it is anticipated that the initial capacity will be 1 000 telephone channels and 4 television channels, the stations being spaced 30 or 40 miles apart, with consequent economy in amplifier station accommodation.

Thus the techniques of radio and land-line communication are merged as the radio links become coupled with the land-line network.

Mr. Twycross is in the Post Office Engineering Department.

DISCUSSION ON 'ELECTRICAL AND MAGNETIC MEASUREMENTS IN AN ELECTRICAL ENGINEERING FACTORY'*

efore the NORTH-EASTERN RADIO AND MEASUREMENT GROUP *at* NEWCASTLE UPON TYNE 31*st November, the* MERSEY AND NORTH WALES CENTRE *at* CHESTER 12*th December*, 1955, *and the* SOUTH MIDLAND CENTRE *at* BIRMINGHAM 5*th March*, 1956.

Mr. C. G. Giles (*at Newcastle upon Tyne*): By the nature of ie processes used in the manufacture of silicon-iron sheets for ectric machines and transformers, wide variations in the losses ay be expected between individual sheets and even within each ieet. For this reason, the usual laboratory measurements have een of doubtful use for obtaining a true mean loss of a batch of ate, because the measurements can of necessity be applied only such a very small proportion of the batch. This is not a true mpling technique, but is the best that could be done under the rcumstances. The design of the whole-plate loss-testing paratus is a considerable achievement, and it should enable the ieet manufacturer to grade every individual sheet to close limits, stead of grading a large batch by the results of tests from ortions of a few sheets selected at random. Thus, by this oser grading, the user should obtain material with much more niform loss and the manufacturer's guarantee will be more presentative of the consignment. The user's check tests in nsequence will be closer to the true mean of the batches tested.

I regret that the author did not show a typical histogram of the ss distribution obtained by the tester on a typical large consign-ient of plate which had been previously graded by Epstein or loyd–Fisher tests. It would have been of considerable interest compare this with similar ones obtained from a large number tests on the Lloyd–Fisher square. Can the author give a pical distribution within a single large sheet?

Power-factor measurements are essential for testing insulating aterials and machine windings. The Carpenter-relay bridge is ovel and economical in apparatus. What is the range of power ctor and sensitivity of the apparatus? In the engineering ctory, speed of testing is of prime importance, and power-factor easurements can be greatly accelerated by the use of an auto-atic dielectric-loss-measuring bridge. This bridge uses a servo stem for balancing, by adjusting the in-phase and quadra-re components of voltage required to neutralize the out-f-balance voltage of the bridge. These are proportional to the ower factor and capacitance changes respectively of the test ecimen. Automatic operation is obtained by servo motors, hich also operate chart recorders, so giving a continuous record power factor and capacitance change of the specimen during e test. The bridge can be arranged to be entirely automatic programming tests, which is a very valuable feature when sting large numbers of similar samples, such as conductor ars, when measurements have to be made at various predeter-ined voltages, which have to be maintained for certain time tervals.

Cdr. J. C. Turnbull (*at Newcastle upon Tyne*): How critical are e values of components in the standard wattmeter amplifier own in Fig. 1, with particular reference to capacitor C and sistor R? What is the order of accuracy of the precision sistors and stable capacitors described in Section 3.2?

Mr. B. Berger (*at Newcastle upon Tyne*): An important class of orks testing is the measurement of the copper loss of large power transformers and current-limiting reactors. Here the power factor may be as low as 0·005, so that conventional measurements may err by as much as 20%. To overcome this difficulty Wilkinson and Carter† developed their toroid bridge in 1939. One difficulty in this bridge is the elimination of stray e.m.f.'s in the toroid and the cable connecting it to the bridge. Another disadvantage is that the bridge is inherently a single-phase tester. In the circuit shown in Fig. 4, not only is the current-transformer phase-angle error unimportant, but it is easily seen that the effect of a small stray signal in the input loop to the amplifier is also unimportant. Whereas three bridge readings are necessary to obtain the total power in a 3-phase circuit, a single measurement will suffice for the method outlined in Fig. 4. In view of these two points, does the author find his circuit more accurate than the toroid bridge?

In dielectric measurements a convenient approach is to measure dispersion and not $\tan\delta$. Dispersion is generally defined as the fractional change in capacitance of a dielectric when tested at two frequencies, say 1 and 100 c/s. It is well known that dispersion and $\tan\delta$ are related quantities, since they both derive from the absorption phenomenon of the dielectric. The dispersion meter presents considerable practical advantages. The model we have designed is portable and battery operated, and hence is convenient for testing on site as well as in the works. Only 1 volt is applied to the dielectric. One may argue that this is an absurdly low test voltage, but if the dielectric is tested in its linear region—and there are many instances in drying and impregnation where one is not interested in void discharges—there is little virtue in using an h.v. bridge. If, on the other hand, one wishes to include the limit of stability, then balance on, say, a conventional Schering bridge is difficult to obtain and the correct approach is to use a self-balancing recording bridge.

Mr. L. C. Kerr (*at Newcastle upon Tyne*): Referring first to the description of the B/H loop tester, it has always been a handicap to the speed of execution of the test when a multi-turn magnetizing coil is wound over the specimen, and considerable time must be saved by using a single turn of heavy cross-section. One problem, however, is that the resistance/inductance ratio of a single-turn magnetizing coil is high, and consequently the voltage waveform is distorted, with a sharp peak. There is therefore a considerably greater initial rate of rise of flux in the specimen from zero up to the knee of the B/H curve than would be obtained with a multi-turn magnetizing coil where the voltage waveform is sinusoidal. From the author's explanation for the differences in values of B_{ac} and B_{dc} given in Table 2 it would appear that the area of the hysteresis loop will be greater below, and smaller above, the knee of the B/H curve with this distorted exciting voltage than for a sinusoidal voltage. Does the author find this in practice, and is there any difference in the total area of the hysteresis loop between distorted and sinusoidal exciting voltages?

Routine permeability tests made in a Lloyd–Fisher square with

* EDMUNDSON, D.: Paper No. 1703 M, September, 1954 (see 102 B, p. 427).

† WILKINSON, K. J. R., and CARTER, G. W.: 'A Method of Measuring Losses in Reactors at low Power Factors', *Beama Journal*, 1939, 44, p. 145.

a.c. excitation must again save considerable time over d.c. testing with permeameters. A further advantage of a.c. excitation is that high flux densities can be induced in the iron for the minimum heat generated in the magnetizing winding, because of the high peak factors of the no-load current waves. Were the instruments shown for current measurements in the circuit of Fig. 13 the only ones used for the a.c. tests when the iron inductions were compared with those obtained with d.c. excitation? At the higher flux densities the author shows that B_{ac} and B_{dc} are in close agreement; here the a.c. no-load current wave has a peak whose amplitude is entirely due to the magnetizing component of current, and no correction need be made to the peak ammeter reading for the iron-loss component before H is calculated. Below about 12kG, however, where the iron permeability is much greater, there is no peak due to the magnetizing component; the no-load current wave is nevertheless still distorted, and only a graphical subtraction of the iron-loss component from the no-load current wave obtained on an oscillograph will give the true magnetizing-current wave and its maximum amplitude. This calculation does not seem possible with the current-reading instruments shown in Fig. 13, and consequently the differences between B_{ac} and B_{dc} may not be so great as those shown in Table 3. Indeed, below 12kG the peak-reading ammeter might well read the peak of the iron-loss current, particularly in tests on the thicker laminations where the eddy-current loss will be some 250% greater than that in the thin plates, and it is perhaps significant that the largest differences are obtained on the thicker plates.

Mr. G. White (*at Newcastle upon Tyne*): The standard wattmeter amplifier is especially useful, since it increases the frequency range and reduces the power consumption of the common dynamometer wattmeter—two very disturbing factors in light-current power measurements. It would have been helpful if the accuracy, stability and reproducibility of the standard wattmeter amplifier had been given.

The frequency-measuring and tachometer circuits incorporate a pointer galvanometer, but presumably they would operate equally well using recorders.

Is the whole-sheet tester for checking the quality of standard-width transformer-steel sheets only, or can it be utilized to check narrower sheets?

Mr. H. S. Chirnside (*at Chester*): About six years ago I had a requirement for testing a large number of small transformer cores, many of them weighing only a fraction of an ounce. They had to be tested for iron loss and magnetizing current, and a number of different ways were tried, but the method finally arrived at for measurement of iron loss used a voltage amplifier not unlike that described in the paper; since then this method has been used for testing many thousands of cores a week. It gives very little trouble and is generally quite satisfactory, but the circuit given in the paper appears to have advantages over the one we use.

Is the whole-sheet tester equally satisfactory at 15kG, the flux density more applicable to cold-rolled transformer sheet?

Mr. J. S. Wilson (*at Chester*): Users of an instrument tend generally to trust its accuracy implicitly. Although many instruments show little change on periodic calibration, frequent checking is a necessity. How much variation has been found in the author's instruments over a period of time, and has it been necessary to check frequently?

In power transformers the core material is used at a fixed flux density. In other spheres the flux density may be considerably lower, and the power to be measured is also small. It would therefore seem advisable to have an amplifier, not only for the voltage measurement, but also for the current measurement. This might also be necessary at high flux densities, to minimize deformation to the flux waveform. Has the author used curren amplification in this or any other sphere of testing?

The contacts of the Carpenter relay are used as a single-pol 2-way switch, and it is possible, by using two such switches in corridor arrangement, to measure differences of speed as deter mined by the current on either side of the corridor. Has th author ever used this arrangement?

Mr. R. D. Haigh (*at Chester*): The fact that the introductio of any measuring instrument must not upset the system bein measured is a fundamental point, but one which is sometime overlooked in industrial measurements, particularly whe checking electronic equipment; earth-leakage measurements ar another example of where very misleading results can be obtaine by the introduction of an instrument.

The author's use of the Carpenter relay as a tachometer o frequency meter is ingenious. I have successfully used a rathe similar principle of charging a condenser and discharging i through a microammeter by means of change-over contacts a a low-speed tachometer, the contacts being operated mechani cally from a multi-lobe cam on the rotating shaft of a flowmeter

It has been my experience that a toroidal sample is the onl satisfactory means of testing the coercive force of soft magneti materials, but I should be glad to know whether the author ha obtained good results by other means, particularly on material having a coercive force of the order of 1 oersted or less.

When plotting B/H loops for magnetizing forces of ± 10 oersteds using direct current with a ballistic galvanometer an mutual inductance, I have noticed differences in the shape of th loop if the current is reversed almost instantaneously by mean of a reversing switch and when it is reversed more slowly, i say $\frac{1}{4}$sec, from positive to negative maximum, by means of a reversing potentiometer. In the former case the knee of th curve is much sharper, although the remanence and coerciv force are the same. Has the author observed this effect an has he any explanation of it?

There is no mention in the paper of the method used fo measuring the insulation resistance (which may be of the orde of thousands of megohms) between two terminals of a com ponent or between a terminal and the case. For such measure ments I have seen used a high-voltage d.c. supply with a mirro galvanometer coupled with a universal shunt, a known hig resistance being used to calibrate the galvanometer. Does th author favour this method, or does he prefer an electronic instru ment which is not so fundamental in principle, but which i certainly more portable?

Mr. H. F. Jones (*at Birmingham*): To what degree of accurac is the author able to grade the large electrical sheets? Hi method for compensating for variations in sheet thickness doe not appear to be particularly good, and if sheets are placed i five grades, considerable overlapping must occur. Moreover, I understand that, when more general use is made of the low-los grain-oriented steels, similar grading will not have to be mad by the user. I should appreciate his further comments on thi point, as I think that some grading is essential, in view of th wide scatter of power lost per pound of this class of material.

Mr. H. K. P. Burt (*at Birmingham*): I feel that the re-entran loop is a challenge to us, and I should like to know whether i is possible that eddy-current losses were included in the calibra tion of the instrument; these losses would normally introduce a vertical ellipse to be added to the B/H loop, and if less durin measurement might produce apparent re-entrance.

What are the methods of ensuring that the flux or the curren in the tests is kept sinusoidal? I think it is important that on or other should be controlled.

Mr. L. G. Ward (*at Birmingham*): My experience with elec tronic meters is that they tend to require rather frequent cali

ation. I am therefore surprised that the author uses a set of em to measure the performance of f.h.p. motors. The motor illustrates is a ¼ h.p. one, and I think that the losses in the eters are quite unimportant in relation to the routine test parently in progress. Is this test performed with electronic eters because of their versatility and self-protecting properties ther than because of their low losses?

Mr. J. Ledbetter (*at Birmingham*): I agree with the author out losses in wattmeters, particularly with regard to measure-ents taken on small motors and transformers. It is an interest-g exercise to take a small 3-phase motor or transformer, run it ht, and measure its losses by the different wattmeter methods. ppreciable differences appear in the results, which must be terpreted with great care.

How stable are the wattmeter amplifiers in operation, and how ten do they require calibration? Once a day has been men-oned; is this necessary or merely precautionary? At what point es saturation begin in the self-protecting circuit, because many ttmeters have overload capacity on both voltage and current nges, and the instrument maintains its accuracy whilst opera-g in this overload region? Are these amplifiers available mmercially?

Has the author any experience of flux measurement by the all effect with germanium? I am interested in the measure-ent of both alternating and direct flux. Is the accuracy greatly ected by temperature? Have any improvements been made to mit the measurement of flux densities of the order a few ndred gauss, such as occur in applications like blow-out coils circuit-breakers and contactors?

Mr. L. S. Moore (*at Birmingham*): The method used for measur-g motor losses is essentially a meter, i.e. a pointer moving over scale, and it takes time for the pointer to settle. Has the author y experience of measuring, say, the stalled power of an f.h.p. otor, where he has perhaps only a second to take the easurement?

Dr. R. D. Gifford (*at Birmingham*): How do the staff at the lling mill deal with the large sheets (8 ft by 3 ft 6 in) coming at the rate of one every 10 sec, and how do they manage to ke the magnetic readings and place the sheets in some sort of der so that they will know how to use them finally?

Mr. A. T. Smart (*at Birmingham*): Can the high-speed relay be ed for long periods at 50 c/s vibration? Does something wear t? Is it important that the adjustment of the contacts is very accurately? Does it matter if the relay does get out adjustment?

Mr. D. Edmundson (*in reply*): Mr. Ledbetter's experience in measurement of low powers agrees with mine, and explains main justification for electronic apparatus queried by Mr. ard, although the other virtues of the amplifier wattmeter umerated in the paper are of considerable importance. Mr. dbetter, like Mr. White and Mr. Wilson, however, queries stability and reliability of the circuit. We find that, as ght be expected, the only components which require occa-nal replacement are the valves, on which it is desirable to ke a monthly check. In general, the stability and repro-cibility of the instrument is better than that of a conventional ttmeter. The ease of performing a daily or weekly calibration ck is a valuable safeguard, but it should be remembered that ulty valve is more likely to give rise to errors of phase angle of magnitude. A calibrating circuit of other than unity r-factor is therefore desirable. The inherent accuracy, ver, is fundamentally dependent on the feedback resistance, h can easily be made much more precise and stable than the ating instrument itself, on which, as Mr. Moore points out, re still dependent. This is the point queried by Cdr. bull, but C is relatively unimportant.

The comparison with Wilkinson and Carter's toroid bridge suggested by Mr. Berger is, as he says, difficult because the latter is a single-phase method. But essentially the limitations are similar, for while the bridge uses a mutual inductance to 'neutralize' the large quadrature voltage, and arrives at the loss component by phase-shifting, the amplifier measures this directly. The sources of error derive from the fact that one is seeking a comparatively small loss component in the presence of a quadrature component a hundred times greater. Phase errors in the quadrature component or stray e.m.f.'s therefore assume great importance. The virtues of the amplifier system are practical: it can be made polyphase, and it is direct-indicating.

Several questions on the subject of the single-sheet tester make it necessary to emphasize that this is not, and was never meant to be, a very versatile instrument. Thus, it could be used on narrower widths, as suggested by Mr. White, but the sheets for which it was designed are nearly always 3 ft wide. Again, it could be used at 15 kG, but not, as Mr. Chirnside inquires, on oriented sheet. This material is delivered in a condition such that a final anneal at the customer's works is required to produce its true magnetic quality. I decided against the inclusion of a histogram, such as Mr. Giles suggests, because it could only refer to production of several years ago. At present all trans-former steel is graded before laboratory samples are taken. The handling problems referred to by Mr. Gifford would certainly worry an electrical manufacturer, but present no difficulties to a steelmaker. Automation is possible, but the method usually favoured is best described as a rock-and-roll process.

Both Mr. Burt and Mr. Kerr point out the difficulty of main-taining a sine wave of either B or H in single-turn specimens. In fact, what is usually controlled is the rate of rise of flux, usually by means of a series reactor. On this depends the loop shape and area. With the type of material usually tested by these means, a sine wave of flux is a practical, and in an ideal sample a theoretical, impossibility. Calibration errors cannot include eddy-current losses; for the B/H tester instruments were calibrated simply as d.c. voltmeters. Mr. Haigh's experience of varying loop shapes is of interest. It seems reasonable to suppose that any loop is a function of the rate of flux change; textbooks are perhaps inadequate on this point.

Mr. Kerr's interesting comments on the a.c. measurement of permeability do not really conflict with the principles suggested in the paper. What we measure is the peak value of B and H in the apparent hysteresis loop. They may not occur at the same instant, but over the range of usefulness usually do. The graphical subtraction which Mr. Kerr suggests will produce a quantity which it would be difficult to define, for the local value of B is often very different from the mean value derived from a search coil. Mr. Wilson's suggestion of an amplifier on the current-measuring circuit is sound. It has been used, and is sometimes essential to avoid distortion. The real difficulty is to make sure that the amplifier is not saturated by the peaky nature of the wave.

The Carpenter relay used in the tachometer is capable of prolonged use, and the contact adjustment can vary widely without inaccuracy. Mr. Wilson's suggestion for a measurement of speed difference, however, would (if I understand him cor-rectly) call for precision adjustment, which is not to be recom-mended. A recorder could be used, as queried by Mr. White, but it would have to be of a servo-operated galvanometer type.

The power-factor tester is used where a dispersion meter—as suggested by Mr. Berger—would hardly apply, i.e. when testing at a range of comparatively high voltages. It is normally used with a range of 0–0·5 in tan δ, but other ranges would be quite practicable.

DISCUSSION ON
'AN INVESTIGATION INTO SOME FUNDAMENTAL PROPERTIES OF STRIP TRANSMISSION LINES WITH THE AID OF AN ELECTROLYTIC TANK'*

Mr. B. G. King (*New Jersey, U.S.A.: communicated*): The accuracy of primary constants of multi-conductor cables measured in an electrolytic tank at the Bell Telephone Laboratories was somewhat better than that recently obtained by Mr. Dukes, and this is partly attributable to differences in technique.

In an electrolytic tank, the potential between electrodes includes that across the capacitance of the monopole layer at the electrode surfaces. Because the inter-electrode potential is assumed to be purely conductive, this reactive potential should be minimized. It is also clear that the distribution of current on the electrodes is affected by this capacitive reactance. Because the potential drop through the layer is not constant, the region of the electrolyte in which ohmic conduction occurs is not bounded by an equipotential surface. Remedially, the frequency can be raised and the current density reduced without reducing the applied potential. We have used a source frequency of 1 500 c/s and an electrolyte with a conductivity of about 10^{-3} mho/m.

The corrosion of electrodes can be mitigated by gold plating and using an inert solute such as boric acid and sodium tetraborate.

The use of a constant-temperature bath surrounding the electrolytic tank and of a vapour barrier over the tank itself prevent changes of electrolyte conductivity due to temperature change and evaporation.

The use of bridge techniques to measure the distribution of current on an element frees the results from the effects of the detector impedance.

The measured and calculated capacitances of a shielded pair made by the Bell Laboratories group agreed to about one part in a thousand over a range of conductor separations from 30% to 70% of the shield diameter. Over a restricted range of conductor separation, the measured attenuation was in essential agreement with the calculated values. It is estimated that the maximum error in the attenuation measurements was 0·5%.

* DUKES, J. M. C.: Paper No. 1991 R, May, 1956 (see 103 B, p. 319).

Mr. J. M. C. Dukes (*in reply*): The improvements to the electrolytic tank described by Mr. King are of considerable practical interest, particularly where a high standard of accuracy is sought; in our own case, however, we were satisfied with $\pm 2\%$ for impedance, and $\pm 8\%$ for attenuation. Nevertheless, the use of a constant-temperature bath, gold-plated electrodes, and an alternative electrolyte would seem to involve very little additional complication and would therefore be desirable in any case. We have no experience of a vapour barrier as such, but we did find it essential to enclose the tank in a dust-proof box (preferably transparent), as this materially reduces both evaporation and contamination. A bridge method was used initially to check the small error introduced by the detector impedance, but was abandoned as it very considerably slowed down the procedure of taking readings. Rapidity of measurement was considered desirable, partly for reasons of convenience and partly to reduce the liability to other sources of error, e.g. evaporation.

In conclusion, there is one general point raised by Mr. King's contribution which deserves brief mention. I have been asked by several colleagues whether for calculations of capacitance, as distinct from attenuation, a tank is necessary, since the capacitance of the actual conductor system can be measured directly with a bridge. In practice, the latter method requires very considerable care to eliminate stray capacitances due to supports, end-effect and adjacent objects. To illustrate this point numerically, consider a conductor system having a capacitance of $10\,\mu\mu$F/ft. Mr. King's technique in effect allows determination of this capacitance to an accuracy of $0{\cdot}01\,\mu\mu$F/ft. With a closed conductor system the difficulty of measuring this small capacitance might be alleviated by the use of a sufficient length of cable (assuming, of course, that the necessary manufacturing tolerance could be met, which is doubtful), but with an open system such as microstrip the difficulties due to stray capacitances would be aggravated.

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